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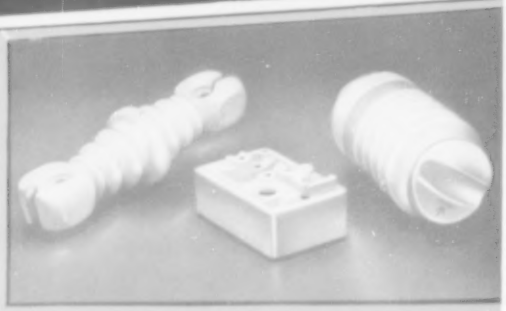

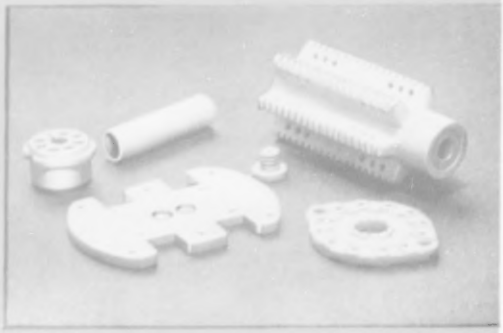

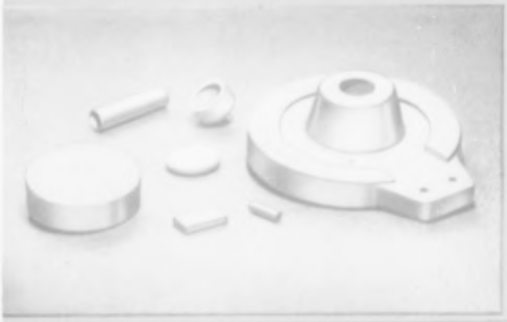

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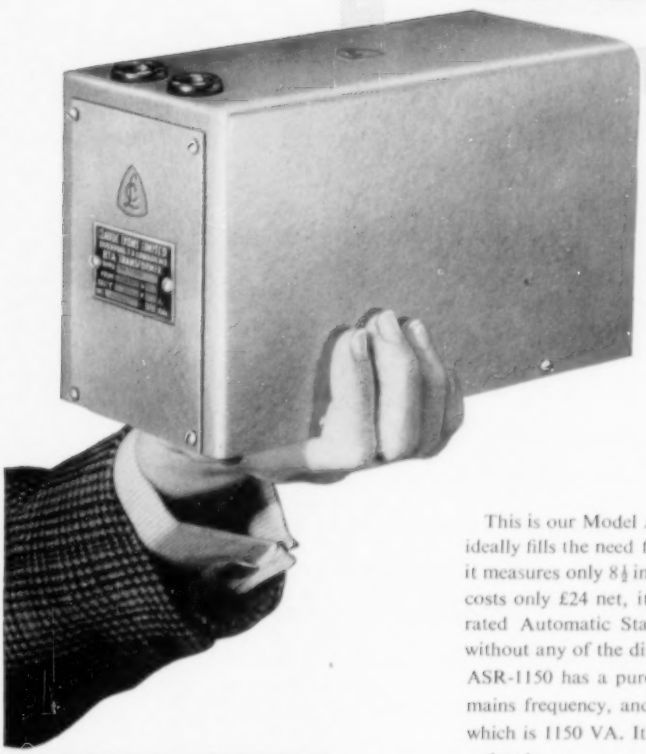

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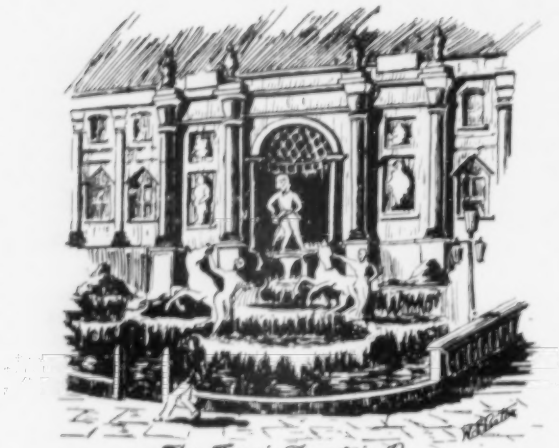

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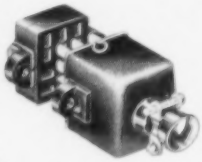

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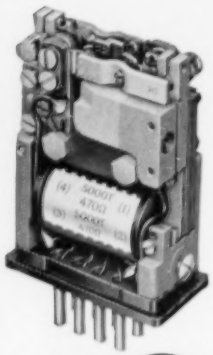

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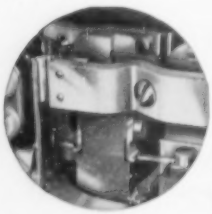

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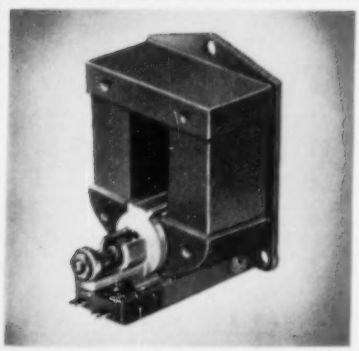

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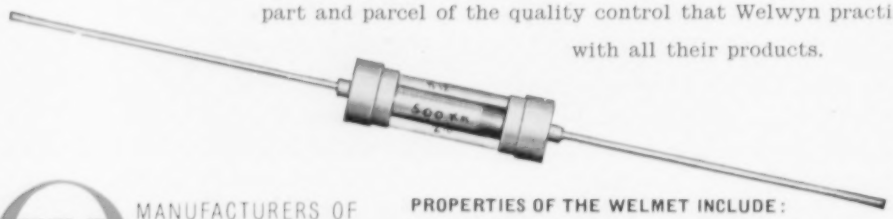

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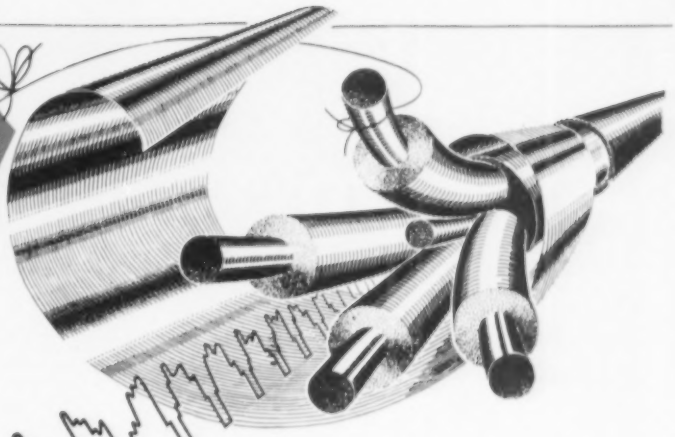

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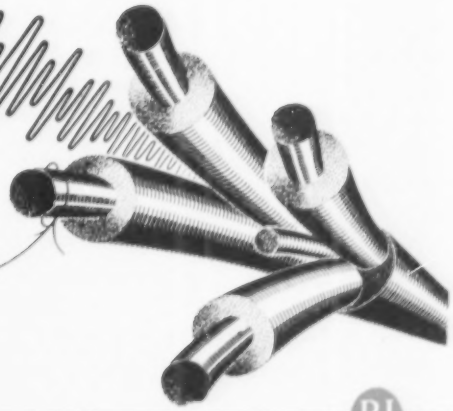

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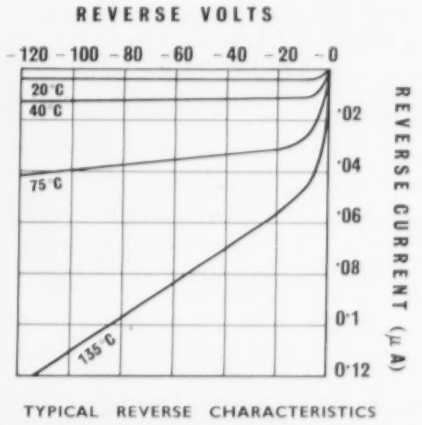

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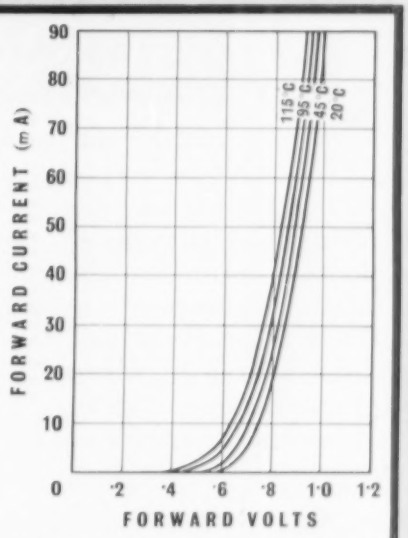

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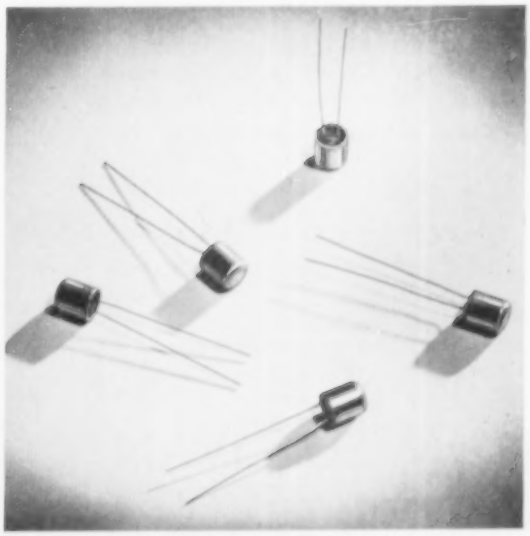

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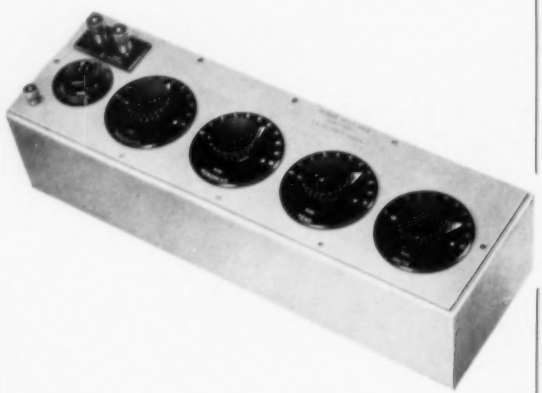

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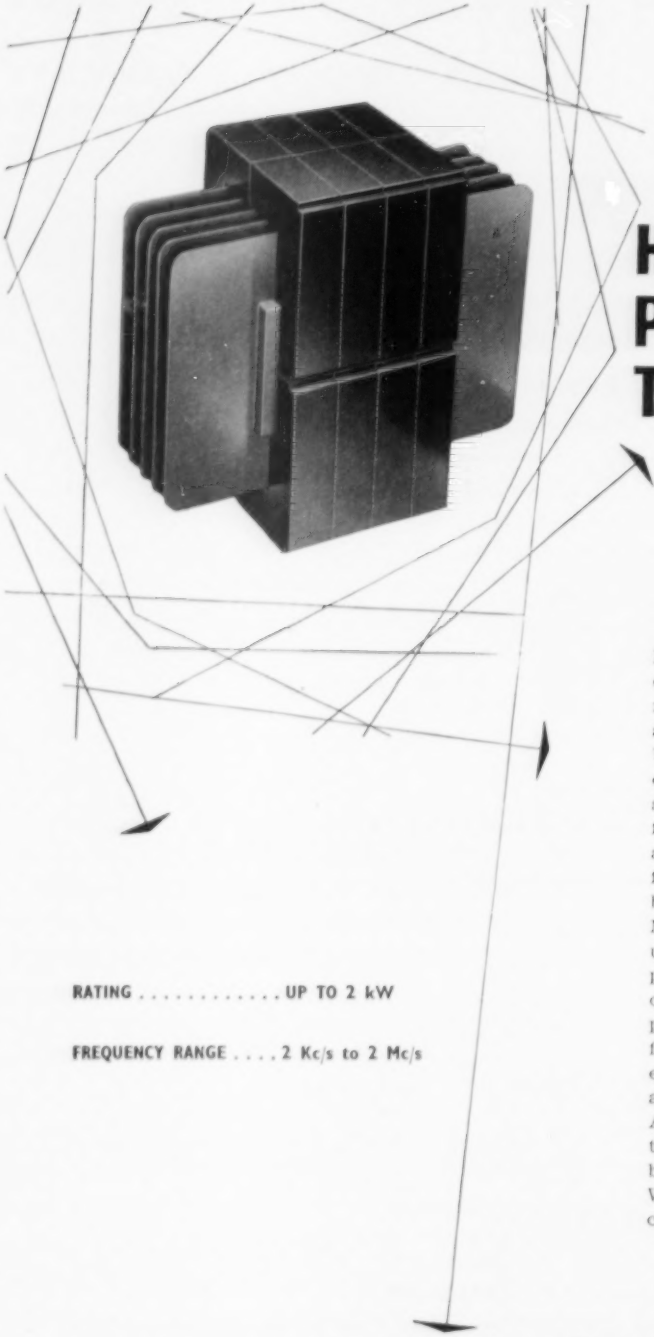

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WIRELESS ENGINEER

Vol. 33 JULY 1956 No. 7

Television Picture Quality

ALTHOUGH there is no immediate prospect of a colour television system coming into operation in this country, there is a good deal of discussion going on about the standards which should be adopted. There is a school of thought which considers that compatibility with the present black-and-white television service is unnecessary and that when colour does come it can be, or even should be, on different standards.

A view which is quite often expressed is that Great Britain should adopt the same colour standards as the Continent and there is good reason to believe that this means a 625-line version of the American N.T.S.C. system. There is, of course, no suggestion that the present 405-line system should be abandoned so far as monochrome transmissions in Bands I and III are concerned. The suggestion is that no attempt should be made to introduce colour into these, but that colour should be developed as a parallel service in Bands IV and V.

Although superficially attractive, there are a good many objections which can be raised against such a scheme. We do not propose to discuss them now, however, for we are more concerned with the underlying idea that 625 lines is a more suitable standard for television than 405 lines. In some circles, it seems to be taken for granted that it is greatly superior. The fact that the Continent only adopted 625 lines after a lengthy examination of our 405-line and the American 525-line systems is often thought to be evidence of its superiority.

We ourselves think that they made a mistake in choosing their standards. We thought so at the time and, having seen 625-line television, we still think so. We do not mean that the 405-line system is better than the 625-line. What we do mean is that the 625-line system does not make the best use of the 5-Mc/s bandwidth allotted to it. We feel strongly that, for this bandwidth, the number of lines should be around 500 and that this would give better pictures than either 405 lines or 625.

The important factors concerned with this matter of lines are not always as clearly realized as they should be. We propose, therefore, to compare them for the two existing systems.

First of all, in the 405-line system, there are 377 active lines; that is, lines conveying visible picture information. In the 625-line system, there are 575 active lines. There is no doubt at all that, in respect of vertical definition, the 625-line system is $575/377 = 1{\cdot}53$ times as good as the 405 line. This improvement of 53% in vertical definition is accompanied by a proportional reduction in the visibility of the line structure, which is in itself a good thing. Too much importance should not be attached to it, however, for methods exist (e.g., spot-wobble) by which the visibility of the line structure can be reduced.

In the horizontal direction, the definition depends on the velocity of the scanning spot and the bandwidth. It is quite usual to assess horizontal definition in terms of a chess-board type of pattern. The fundamental component of the waveform produced by scanning such a pattern is a sinusoid of which one cycle corresponds to two adjacent elements of the chess-board. If the duration of the active part of a line is τ_1 and there are n elements per line, the duration of one cycle of this sine wave is $2\tau_1/n$.

The frequency is thus $n/2\tau_1$ and this is taken as the upper limit of the video bandwidth.

This is rather an arbitrary relation, but it is practically useful and, in any case, it is valid for comparing different systems. With 405 lines, the frequency is 3 Mc/s and τ_1 is 80·95 μsec. Therefore, $n = 2 \times 3 \times 80{\cdot}95 = 485{\cdot}7$ elements per line. With 625 lines, the frequency used is 5 Mc/s and τ_1 is 52·32 μsec, so $n = 2 \times 5 \times 52{\cdot}32 = 523{\cdot}2$ elements per line. The horizontal definition is thus $523{\cdot}2/485{\cdot}7 = 1{\cdot}075$ times that of the 405 line; that is, an improvement of 7·5%.

The facts are thus, that 625 lines gives 53% better vertical and 7·5% better horizontal definition than 405 lines. For this, it requires a greater bandwidth. The video bandwidth is greater in the ratio of $5/3 = 1{\cdot}66$ and the channel bandwidth (i.e., the band occupied by both sound and vision transmissions) in the ratio $7/5 = 1{\cdot}4$.

The aspect ratio is 4/3 in both cases. With 405 lines, there are $485{\cdot}7 \times 3/4 = 364$ elements in the same linear distance as the 377 active lines. The definition is thus virtually the same in both the vertical and horizontal directions. With 625 lines, there are $523{\cdot}2 \times 3/4 = 392$ elements in the same linear distance as the 575 active lines. The vertical definition is thus $575/392 = 1{\cdot}46$ times as good as the horizontal.

It is clear that the 625-line system gives better definition than the 405 line but at the expense of 40% more channel width. The improvement is almost entirely in vertical definition. For practical purposes, the two are virtually the same in horizontal definition.

If the bandwidth is kept constant, altering the number of lines improves the definition in one direction across the picture at the expense of the other. It is clear that there must be an optimum number of lines and it is not unreasonable to suppose that it occurs when the definitions in the two directions are the same, as in the 405-line system. If the two are not the same, in which direction should the definition be the better? We ourselves feel that horizontal definition is more important.

In support of this view, we instance the following:—

(a) In the cinema, the tendency is all to wide screens, emphasizing the horizontal direction.

(b) In laboratory tests of 405-line and 625-line pictures, each having the bandwidth of its particular standard, we find it hard to say with certainty which is which, unless there is a direct comparison.

(c) In laboratory tests of the 405-line system with unrestricted bandwidth against 625 lines with 5-Mc/s bandwidth, we find that the 405-line system gives the better picture.

In the 625-line system, the extra 2 Mc/s of bandwidth is used almost entirely to increase the vertical definition. It is this that we think wrong. We think that 405 lines with a 5-Mc/s bandwidth would give a better picture. We do not say, however, that the proper number of lines for this bandwidth is 405. Probably the right thing to do is to keep the definition in both directions the same and, for 5-Mc/s bandwidth, that means about 500 lines, as we said earlier.

It should be unnecessary but, in view of the figure of 500, perhaps we should point out that this does not mean that we think the American 525-line standard is the right one. That has a 60-c/s frame frequency and a 4-Mc/s bandwidth and gives inferior horizontal definition to the 405-line system.

In saying all this, we are not advocating any change of standards by anyone. So far as monochrome television is concerned, that is impracticable. What we are concerned about is that, if the introduction of colour is made on standards other than 405 lines, the new standards should be the right ones. If it is practicable to have different standards for colour, we do not think that 625 lines with 5-Mc/s bandwidth is right. For this bandwidth, 500 lines or so would be better but, better still, would be 625 lines with the 7·4-Mc/s bandwidth necessary to equalize the definition in the two directions.

W. T. C.

OBLIQUE TRANSMISSION BY THE METEORIC E-LAYER

By R. Naismith, M.I.E.E.

(Official communication from D.S.I.R. Radio Research Station, Slough)

SUMMARY.—This paper describes the results of observations made on transmissions between England and Norway on frequencies of 22·7, 25 and 27 Mc/s over distances of 900 to 1,900 km. The observations, which were made in the winter of 1951 and the summer of 1952, indicate that propagation took place by way of that part of the ionosphere which has been designated the meteoric E-layer. Measurements of the conditions in this layer made at vertical incidence near one end of the transmission path provide some confirmation that this mode of propagation was effective during the period of the experiments. It is concluded that this meteoric E-layer may enable communication to be maintained over certain distances at frequencies higher than those normally propagated by way of the E and F layers.

THE term meteoric E-layer refers to a layer situated in the lower part of the ionosphere which is observable for about 85% of the time. It is an inferior reflector of radio waves. The following observations were planned to confirm the suggestion[1] that it would enable higher frequencies than those normally reflected from the ionosphere to be received at distances up to the limit of a first reflection (i.e., about 2,000 km) when high-power transmitters were used. Suitable sources of such high power are the radar stations along the east coast of England, some of which had already been used by Eastwood and Mercer[2] to demonstrate that individual echoes could be observed back at the transmitter after reflection at oblique incidence. They further concluded that the reflections they observed originated in a thin layer situated at a uniform height which we may now associate with the meteoric E-layer. Since, however, the limit of resolution of their equipment was reached when the echo duration was less than 0·3 second, the present observations may also be regarded as extending their work to determine if the integrated effect of the many shorter duration echoes would result in a quasi-continuous signal over an oblique path. In this case it would be necessary to make the observations abroad to ensure that the receivers were out of range of the direct ray from the radar stations.

An arrangement was therefore made between the British Air Ministry and the Norwegian Defence Research Establishment which enabled observations on the radar stations to be made in Norway during a winter period in 1951 and a summer period in 1952. The orientation of the radar beam and the power used were unknown so far as this experiment is concerned and no effort was made to extend the observations by measuring the strength of the signals. It was convenient to use an AR88 type of receiver although its comparatively narrow bandwidth was not very suitable for the reception of radar signals. If it were possible to receive the signals under such unfavourable conditions the suggestion would be proved correct.

The frequencies used were 22·7, 25 and 27 Mc/s and the distances varied from 900 to 1,900 km. The observing stations were near Oslo, Trondheim and Bardufosse. The latter, situated at latitude 70°N, was the farthest north at which radar transmissions were observed. Since the Norwegian mountains were in the path of any direct signals it is quite certain that the signals received at Trondheim and Bardufosse travelled by reflection in the ionosphere. Also, by receiving different radar stations at one place it proved unnecessary to be directly in the line of maximum radiation.

An unsuccessful effort was made to receive the signals at latitude 78°N. It is possible that the distance, which was over 3,000 km, was too great to be covered in a single hop and the low reflection coefficient made it impossible to receive a twice-reflected signal. Observations were also unsuccessful over a ground distance of 600–700 km and this may be attributed to a skip effect. Thus, the range of distances over which the signals were received conforms to the idea of a single reflection in the ionosphere.

The observations were made on the normal radar transmissions and covered a total of 1,000 hours (445 in winter and 555 in summer). They were made by observing each frequency in turn and not by continuously monitoring on a single frequency. The latter would have shown a longer period of reception since the comparatively weak signals were, at times, difficult to separate from the general noise. The figures in the Table show the percentage of the total number of hours of transmission during which the stations were received, and indicate that for 33–54% of these hours it was possible to record radar signals over distances of about 1,000 km in both summer and winter.

MS accepted by the Editor, July 1955

The number of observations is too small to enable the variations with frequency, season, distance and latitude to be assessed but the percentages shown would, of course, be improved upon if complete control of the transmitters and receivers were possible. The observations show quite definitely that radar signals intended for reception only within the range of direct transmission may be received at distances up to 2,000 km and at places well within the Arctic circle even in midwinter.

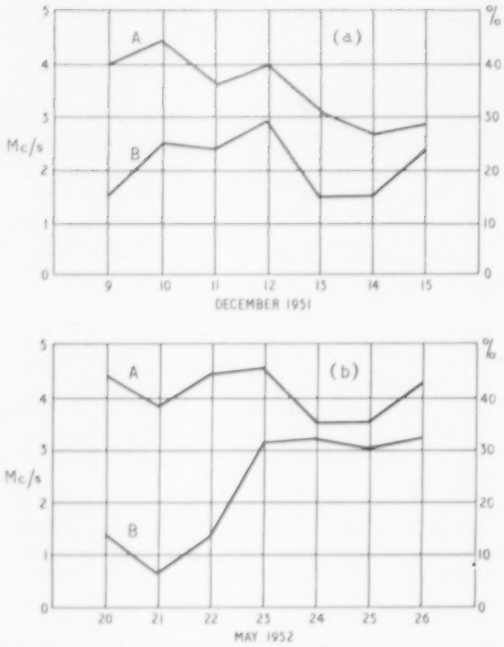

Fig. 1. *Results for December (a) and May (b). In each case, curve A is the daily average of fEM and B is the daily percentage reception in hours.*

Conditions of Meteoric E-Layer

Since the suggested experiment arose from a study of the characteristics of the meteoric E-layer it is worth while setting down the conditions in that layer as they were observed at Slough during the progress of the experiment.

The hourly values of the highest frequency on which echoes were received from the meteoric E-layer at vertical incidence (fEM) have been averaged over the day and plotted for each day of the tests in Fig. 1 (a) and (b). The percentage of the hours of reception of the radar transmissions has been treated similarly and plotted on the same graphs. During three days (20th–22nd May) absorption obscured all but a few E-region echoes at vertical incidence at Tromso, which is situated near Bardufosse. Taking this into account there appears to be a general correspondence between the average values of fEM and the average number of hours of reception. For 75% of the days the values of fEM observed at Slough increased and decreased as the number of hours of reception of the radar signals in Norway increased and decreased. For the remaining 25% of the time the values are not in opposition. There are insufficient observations to make a statistical interpretation significant; it is necessary, therefore, to leave the observations in the above form. This shows that meteoric E-echoes were present during the tests and that, with a sufficiently powerful transmitter, reception would be expected to occur in the circumstances of the experiment as described.

Turbulence

There is little doubt that a certain amount of turbulence is present in all regions of the ionosphere. Indeed, even if no other source of turbulence were present, the incidence of sporadic and shower meteors would cause it. The large amount of local heating and the high peak values of ionization produced are quite certain to produce turbulence effects. So far as I know only one experiment has been performed which might have been expected to demonstrate the effect of turbulence. On 9th July 1945 a solar eclipse occurred in which the totality belt passed near to the north of Scotland. A chain of radar stations which had already reported the frequent reception of meteor bursts was suitably situated for observation of the occurrence of these bursts in this region of totality and I suggested that this should be done to determine, if possible, whether the eclipse of the sun would affect the burst formation.

This experiment was carried out under R.A.F. auspices and the results carefully analysed by

TABLE

Station	December 1951			May 1952			Approx. Distance (km)
	22·7 Mc/s	25·1 Mc/s	27 Mc/s	22·7 Mc/s	25·1 Mc/s	27 Mc/s	
Oslo	17	25	38	7·5	50	33	900
Trondheim	54	0	0	33	29	26	1,100
Bardufosse	13	8·5	2·5	17	41	21	1,900

Eastwood and Mercer who concluded[2] that the burst activity throughout the eclipse period failed to reveal any significant departure from the averages established on the two control periods each of seven days' duration, on either side of the eclipse day.

As another part of the same experiment we measured the critical frequency for the E-region (fE) at Loth in the north of Scotland where the sun was 75% obscured at the maximum phase of the eclipse. A drop of 40% in the ionization in this region occurred. This made the value of fE at 1430 LMT (local mean time) on the eclipse day, correspond to 1800 LMT on a normal day. This corresponds roughly to a difference of 8 dB in the field strength reported for June 1951 by Bailey et al[3]. Actually, the reduction would be much greater than this in the totality belt and some of the meteor observations were made in or near to that region. The conclusions reached by Eastwood and Mercer are therefore most significant from the turbulence aspect of this problem.

Again, if turbulence were a predominant characteristic of the ionosphere one might expect to see some effects in the F_2 region particularly on a winter morning in the northern hemisphere when both the amount of ionization present and the rate of increase should give rise to favourable conditions for the creation of turbulence effects. The absence of any confirmatory reports of turbulence effects in the F_2 region is therefore surprising. We also know that there are considerable movements observable—sometimes at high speeds—on frequencies reflected from the F_2 region in the more normal manner.

On the other hand, the meteor-type of signal has been observed on a winter night at latitude 70°N when only a vestigial amount of ionization remained in the normal E-region and when, therefore, the turbulence effect seemed unlikely.

Further simultaneous observations on the meteoric E-region at oblique incidence to the north of Aberdeen and to the south of Weymouth have shown no systematic difference in the number of echoes observed due to sporadic meteors, although the scattering centres would be about 1,000 km apart. The amount of ionization in the region in which turbulence could occur was much greater at the more southerly observing site.

This uniformity over large areas may well be a characteristic of observations on sporadic meteors. Lovell[4] estimated the number of sporadic meteors arriving in the earth's atmosphere as 8×10^9 over 24 hours and their limiting velocity as 72 km/sec. Thus meteor dust particles take only one or two minutes to travel the distance from the United Kingdom to the United States and it is unlikely, therefore, that there will be any significant difference in the amount of this dust arriving over large areas of the world. We may relate this suggestion to the following observation on a frequency higher than the maximum usable for either regions E or F.

Monthly average values of signal strength received on 49·8 Mc/s over a distance of 1,245 km in America have been published for the months of April and June 1951. The authors remark that "some preliminary speculations suggest that the mechanism of this type of propagation may be scattering caused by the ever-present irregularities in the E-region and an appropriate transmission equation is derived in terms of parameters describing inhomogeneities in the E-region". However, they do not rule out the possibility of a contribution made by meteors which could explain some of the characteristics of their observation. If, however, sporadic meteor dust is mainly responsible for this type of radio propagation then we may expect to observe similar monthly average conditions over the United States and over the United Kingdom. On the other hand, we know that scattering arising from turbulence—as observed in tropospheric propagation—is much less likely to give the same monthly average conditions in these two areas and this should also apply in the case of turbulence observed in ionospheric propagation. If it can be shown that a close similarity does, in fact, exist between the fEM measurements at Slough and the American observations referred to above, additional evidence would be deduced for attributing the observations of the reception of British radar signals in Norway to the incidence of sporadic meteors rather than to turbulence.

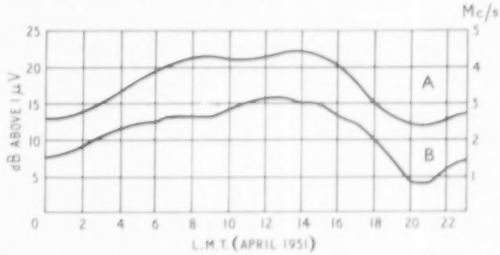

Fig. 2. *Curve A shows the monthly average fEM and B the monthly average signal intensity on* 49·8 *Mc/s over a distance of* 1,245 *km.*

In Fig. 2 the American observations and the average values of fEM measured at Slough have been plotted. The similarity of the two graphs is apparent. The pre-sunrise minimum strongly suggests a meteoric origin as the source of the ionization involved in the reflecting process. A comparison of corresponding data for June 1951 gives similar results.

Conclusion

The suggestion that the meteoric E-layer enables a type of long-distance radio communication on frequencies higher than those normally reflected from regions E and F has been demonstrated. Additional confirmation of the meteoric source of the ionization involved has been deduced from earlier observations of the phenomenon of this type of radio propagation at very high frequencies.

Acknowledgments

Thanks are due to Mr. C. F. Sutton of Signal Plans, Air Ministry, and to Mr. S. Skribeland of the Norwegian Defence Research Establishment for the care taken in arranging for reliable measurements to be made on the routine radar transmissions.

The work described above was carried out as part of the programme of the Radio Research Board. This paper is published by permission of the Director of Radio Research of the Department of Scientific and Industrial Research.

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KLYSTRON CONTROL SYSTEM

By R. J. D. Reeves

(E. K. Cole, Ltd., Malmesbury, Wilts)

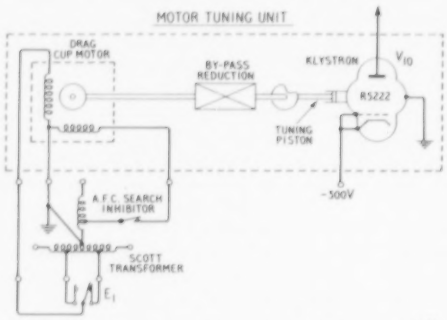
(Continued from p. 143, June issue)

Part 3—Specimen Design for A.F.C.

General

FIG. 13 shows a complete circuit diagram for an A.F.C. system embodying the features described in Part 2. It was designed for a radar with a $\frac{1}{2}$ μsec pulse and a repetition rate of 2 kc/s, and the period of the sinusoid was chosen to be 20 μsecs. The local oscillator V10 is a R5222, which has an external resonant cavity and therefore affords full freedom of mechanical design. The frequency sweep is about 500 Mc/s in the X band (3 cm). The motor tuning unit is shown Fig. 14(a), and the rest of the control system is divided into two parts; the frequency discriminator unit Fig. 14(b), and the klystron control unit Fig. 14(c).

A negative supply of 300 V is necessary for the klystron and is freely employed elsewhere in the circuit. The positive supply is 150 V which is an economically low value and conveniently



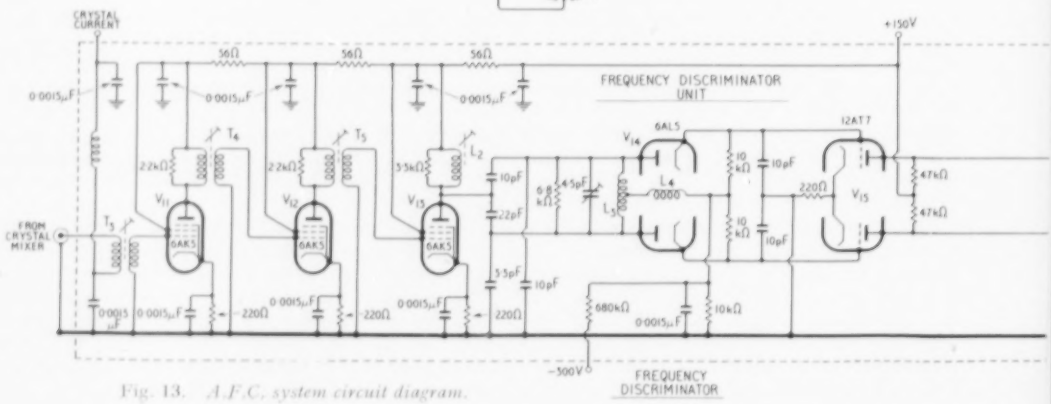
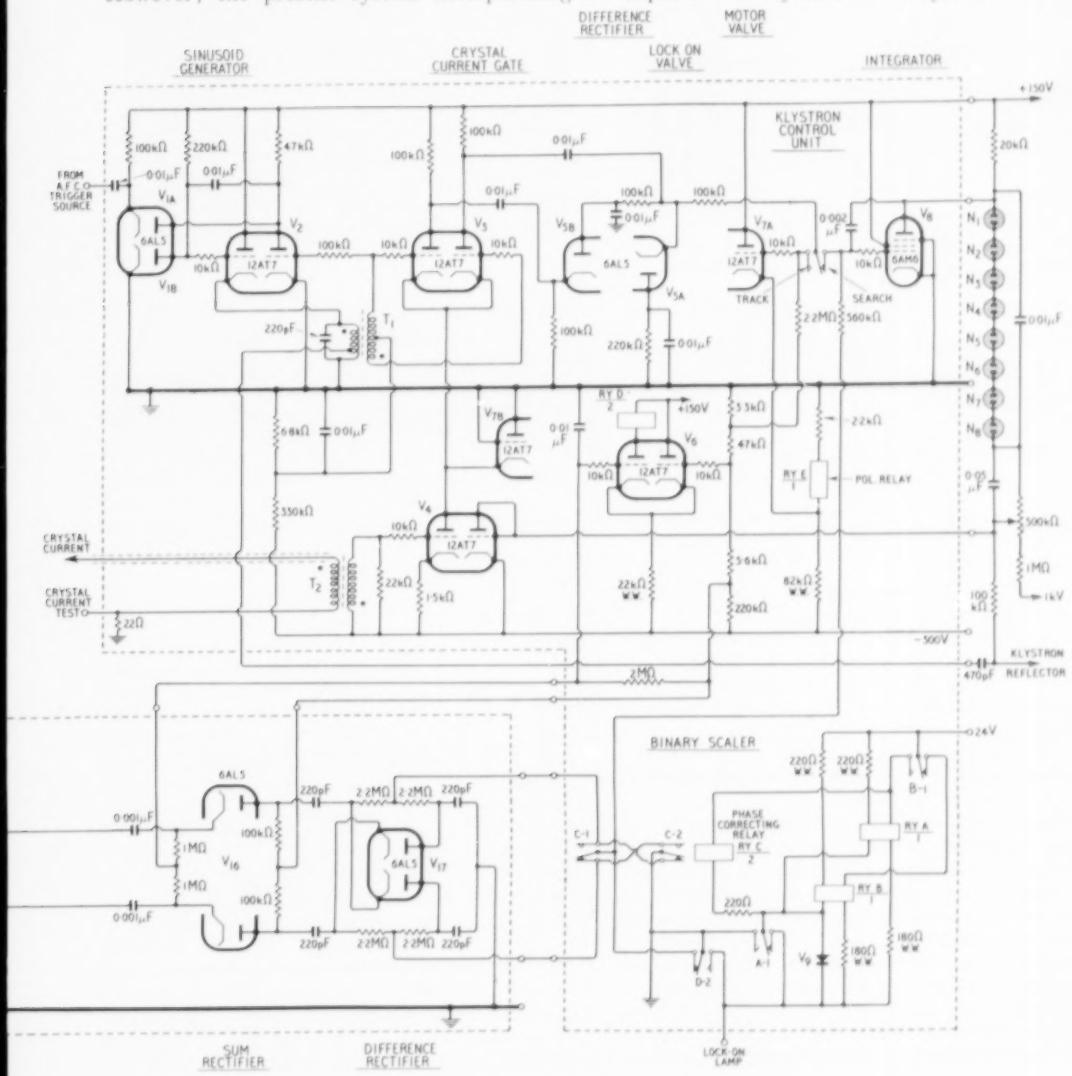


Fig. 13. *A.F.C. system circuit diagram.*

limits the anode base of the Miller valve V_8. A twenty-four volt line is introduced to operate the relays of the binary scaler, and the potentiometer to the klystron reflector is returned to a —1 kV source which is normally available in a transmitter for the 'keep alive' electrodes of the t.r. cells. It has been suggested that an r.f. oscillator and detector be used as a link between the Miller anode and the klystron reflector in order to obviate the need for this high voltage supply, and it would certainly be an attractive possibility if the —1-kV line were not already available. However, the present system incorporating a chain of miniature neons has the useful property of almost completely relieving the control system of the onus of dealing with errors due to supply-voltage drifts for, by holding the appropriate fraction of the reflector potential constant, the residual variations appearing on all the electrodes of the klystron combine to have approximately zero effect on the frequency being generated. For this reason the neon chain is to be preferred to other methods.

In the following sections the individual circuit functions are discussed in detail and certain aspects of the system are analysed.



The Sinusoid Generator

Fig. 15 is the section of the circuit which produces one cycle of a sine wave when triggered by a negative pulse. One winding of a tuned transformer T_1 carries the current of the triode V_{2B}, and the energy stored causes the tuned circuit to ring when the valve is shut off. Neglecting losses, the peak amplitude of the ring is:

$$V = I\sqrt{L/C} \text{ volts} \quad \ldots \quad \ldots \quad \ldots \quad (1)$$

where I is the anode current of the valve when conducting, and C is the capacitance which tunes the current-carrying inductance L.

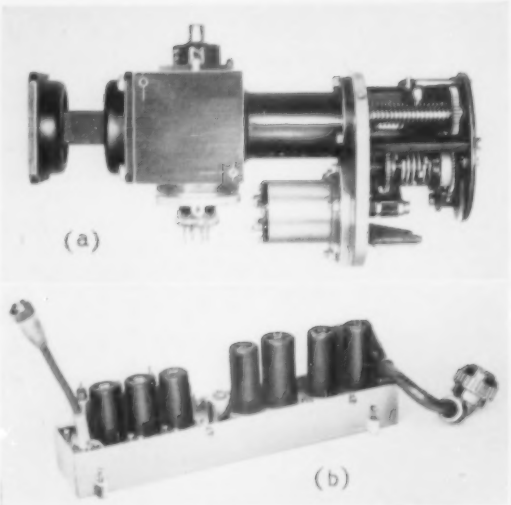
Initially the shut-off is achieved by a trigger pulse passing through the diode V_{1A} and the a.c. coupling to the ringing-valve grid, but the condition is sustained by coupling the ringing waveform itself back to this grid through the saturating amplifier V_{2A}. Being fully conductive, this valve holds V_{2B} cut-off during the first half cycle of the sinusoid. In the next half cycle V_{2A} is switched off and V_{2B} grid recovers towards a positive potential until it is caught at earth by the diode V_{1B}. There is still no current in the ringing valve, however, because its cathode has run positive with the resonant circuit, and it remains shut off until the end of the second half cycle, when the cathode swings into the grid base. Then the valve abruptly damps out the oscillation.

Because the sinusoid itself performs the switching functions, the shut-off time is bound to correspond precisely to one ringing cycle and there is little that is critical about this circuit. The waveform that appears on the grid of V_{2B} is shown in relation to the cathode swing in the oscillogram of Fig. 16. The slow recovery of the grid waveform is necessary to effect the transition from the first to the second half cycles of the ring.

The circuit is prevented from free running by a definite delay bias on V_{2A} grid.

Time Discriminator

A fraction of the ringing waveform is picked off from T_1 and fed to the reflector electrode of the klystron. The resulting variations of oscillation strength in that valve, as observed by the current in a crystal mixer, are fed through the pulse transformer T_2 to the video input valve V_4 of the gating circuit (see Fig. 17). The anode current of this valve normally flows through the diode V_{7B} since the grids of V_3 are returned through the secondary of T_1 to a cut-off bias. However, when the sinusoid is generated, first one grid and then the other will swing well



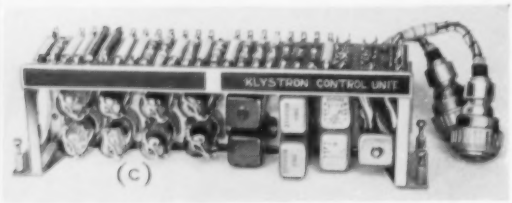


Fig. 14. *Component units: (a) Tuning mechanism; (b) Frequency discriminator unit; (c) Klystron control unit.*

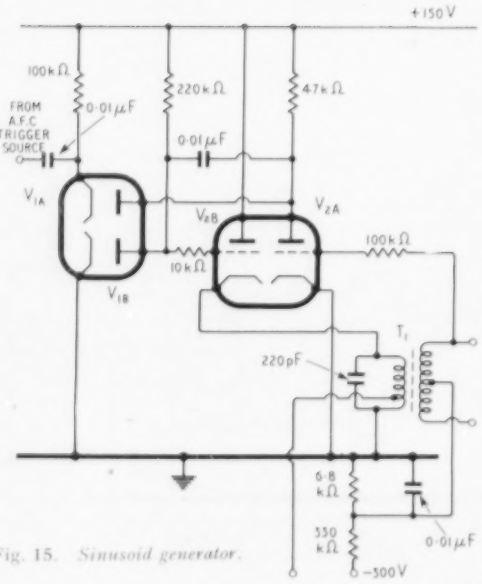


Fig. 15. *Sinusoid generator.*

positive with respect to earth and divert the anode current to the valve with the positive grid. In this way the modulated crystal current is separated into two channels which can be identified either as the 'early and late' or the 'negative and positive' channels, since they are associated with the first and second half periods of the sinusoid. The current pulses that pass through the gate valves are integrated in capacitors, and the

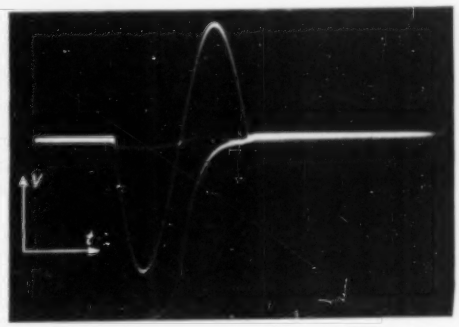


Fig. 16. *Grid and cathode waveforms of ringing valve.*

charge differences build up an error potential (see next section).

The features of this circuit to notice are, first, that the gate current is largely independent of the gate amplitude, being defined by the video input to the 'tail' valve V_4, and secondly, the gate generator is not loaded since the cathodes of V_3 follow the positive grid. Good switching action is obtained if the sinusoid amplitude at the transformer secondary is large compared with the standing bias.

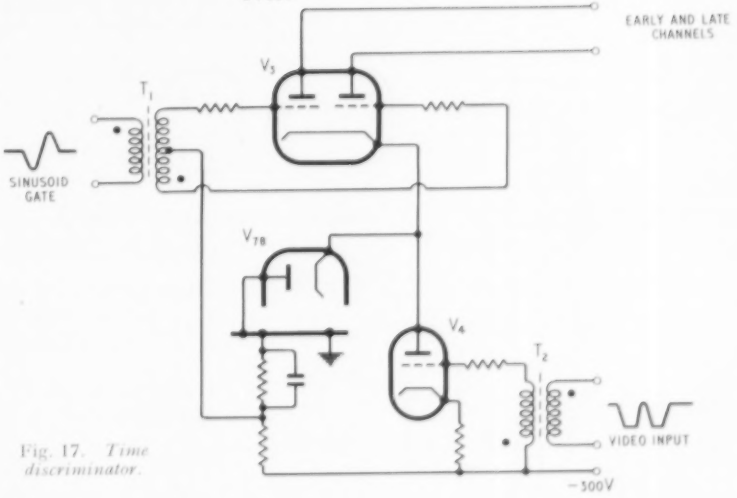


Fig. 17. *Time discriminator.*

The Difference Rectifiers

The function of both the time and frequency discriminators is to produce a steady or 'clamped' error signal with magnitude and sign related to that of the misalignment measured at the sampling instants. This task is simplified if a measure of integration is permissible; that is to say, if the error output may be built up from not just the last sample but all previous samples as a weighted aggregate. The discriminators are not then the traditional error-clamp generators of popular hypothesis, but are inseparable from either a time constant or a true integration effect. The error output circuits of both the discriminators are derived from that of Fig. 18. At the sampling times a current pulse is passed by the early and late gate valves and it is the difference in the area under these two pulses that constitutes the error signal. Only a small fraction of the input current is required to develop sufficient voltage across the resistors to switch on the diodes and, when these are conducting, the impedance in each gate-valve anode is just that of C_1 and C_2 in series. A charge then accumulates on these capacitors with a polarity which will hold off the diodes when the charging current ends. Until the next charging time there are therefore two independent but identical meshes comprising C_1, R_1, C_2, R_2 in series, and the capacitors in a common mesh are able to discharge into each other towards a final potential representing the charge difference. It may be demonstrated that if the CR product is the same each side of the pick-off junction then the relaxation waveform is absent at this point and the final potential is available immediately the charging interval is over.

In the circuit of Fig. 18 this output is available in balanced form from the two meshes, but this configuration is severely limited in the magnitude of error signal it can support when the diodes are biased entirely by the relaxation waveform on their anodes. A practical form of the circuit is shown in Fig. 19 in which no attempt is made to accumulate an error signal in both relaxation meshes but only in the upper one. The lower mesh is split into separate parts by an earth junction and the relaxation potentials on nodes

(5) and (6) [with that of node (3)], serve to bias the diodes without mutual interference. In fact, the components of the lower mesh are not related to the output circuit and the biasing waveforms can be freely designed to support a substantial error signal at node (2). This configuration is the one employed for the mode-centre discriminator.

In the pulsed frequency discriminator there is a special requirement that the output shall be available in either polarity (although not both simultaneously) and for this reason a symmetric

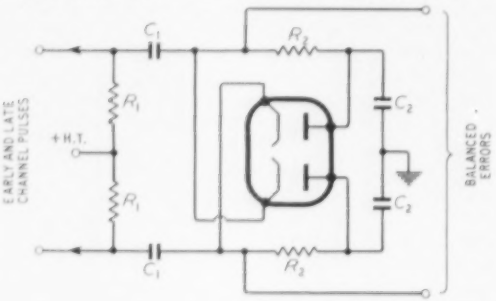


Fig. 18 (above). *Basic difference rectifier circuit.*

Fig. 19 (below). *Practical form of difference circuit.*

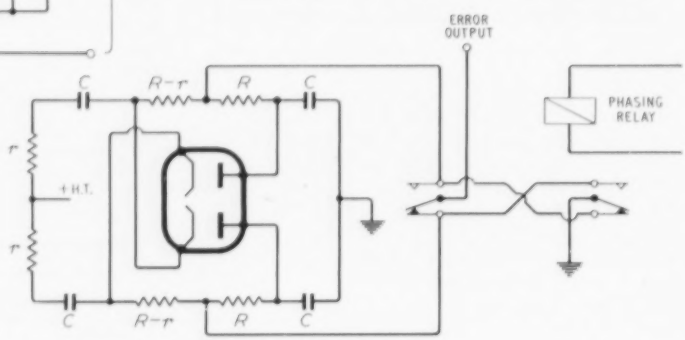
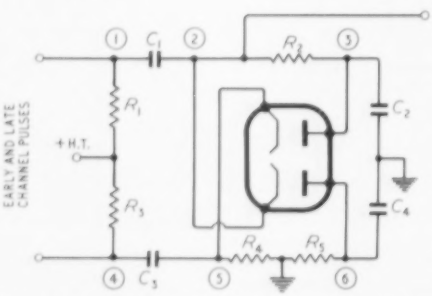


Fig. 20 (right). *Balanced form of difference circuit.*



form is reverted to (see Fig. 20). The upper and lower meshes are identical, with pick-off junctions at the balance point of each.

These junctions are not coincident with the diode cathodes and so either may be used as an earth point without interfering with the charging paths. By using a two-pole change-over relay which selects one junction and earths the other an error signal of either polarity may be extracted.

Transfer Function of the Difference Rectifiers

The transfer function of the frequency discriminator is of fundamental importance when investigating the limitation of the complete a.f.c. system, whereas that of the time discriminator is of secondary interest. It is informative, however, to examine the behaviour of the most elementary difference rectifier circuits, such as the one employed in the time discriminator.

The upper relaxation network is drawn in Fig. 21 and by inspection its nodal determinant is:

$$\Delta = \begin{vmatrix} \frac{1}{R_1} + C_1 p & -C_1 p & . \\ -C_1 p & \frac{1}{R_2} + C_1 p & -\frac{1}{R_2} \\ . & -\frac{1}{R_2} & \frac{1}{R_2} + C_2 p \end{vmatrix} \qquad (2)$$

The voltage on node (2) is given by:

$$\frac{(\Delta_{22} + \Delta_{12})I_A + \Delta_{32} I_B}{\Delta}$$

which gives:

$$E_2 = \frac{(1 + C_2 R_2 p)I_A - (1 + C_1 R_1 p)I_B}{(1 + C_2 R_2 p)C_1 p + (1 + C_1 R_1 p)C_2 p} \qquad \ldots \quad \ldots \qquad (3)$$

It will be observed that by putting $C_1 R_1 = C_2 R_2$ this reduces to

$$E_2 = \frac{I_A - I_B}{(C_1 + C_2)p} \qquad \ldots \quad \ldots \quad \ldots \qquad (4)$$

That is to say, the voltage on node (2) is the integral of the current difference, so that immediately after simultaneous current impulses of magnitude Q_A and Q_B there is a potential difference

$$e_2 = \frac{Q_A - Q_B}{C_1 + C_2} \qquad \ldots \quad \ldots \quad \ldots \qquad (5)$$

without extraneous relaxation waveforms. (There will be found unwanted impulses at the charging instants but these are more easily filtered off).

This simple configuration can therefore be

considered to be a high-speed error clamp which integrates the input—provided it is not loaded. If it is loaded with resistance, there will be a relaxation waveform and a weighted memory, whereas if there is capacitance loading there will only be a relaxation waveform. The circuit of the frequency discriminator is of this type and will be examined in more detail.

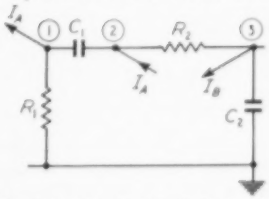


Fig. 21. *Output mesh of Fig. 19.*

Fig. 22 represents the relaxation mesh of a difference rectifier loaded with a Miller integrator, which corresponds to the output circuit of our frequency discriminator. The symbol S denotes the mutual conductance of V_8.

The nodal determinant of the circuit is

$$\Delta = \begin{vmatrix} Cp+\frac{1}{r} & -Cp & . & . & . \\ -Cp & Cp+\frac{1}{R-r}+\frac{1}{R-r} & -\frac{1}{R-r} & . & . \\ . & -\frac{1}{R-r} & kCp+\frac{1}{R}+\frac{1}{R-r} & -\frac{1}{R} & -kCp \\ . & . & -\frac{1}{R} & Cp+\frac{1}{R} & . \\ . & . & S-kCp & . & kCp \end{vmatrix} \quad (6)$$

and the output voltage is given by:

$$\begin{aligned} E_5 &= -\frac{(\Delta_{15}+\Delta_{25})}{\Delta} I_A + \frac{\Delta_{45}}{\Delta} I_B \\ &= \frac{(S-kCp)(I_B-I_A)}{kCp[2Cp+S(CRp+1)]} \\ &\approx \frac{(S-kCp)(I_B-I_A)}{kCpS(CRp+1)} \quad (\text{since } SR \gg 2) \\ &= \frac{1}{kCp}(I_B-I_A) - \left(\frac{1}{SR}+\frac{1}{k}\right)\frac{1}{C}\frac{I_B-I_A}{\left(p+\frac{1}{CR}\right)} \end{aligned} \quad \ldots \quad \ldots \quad (7)$$

$$\therefore e_5(t) = \int \frac{I_B-I_A}{kC}\,dt - \left(\frac{1}{SR}+\frac{1}{k}\right)\frac{(I_B-I_A)}{C}\epsilon^{-t/CR} \quad \ldots \quad \ldots \quad (8)$$

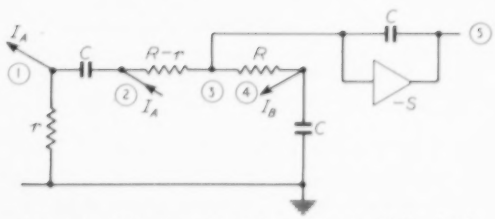


Fig. 22. *Output mesh of Fig. 20 with Miller load.*

The response to impulses of magnitude Q_A and Q_B is shown in Fig. 23. The final voltage is determined by the Miller capacitance kC and the time taken to reach it is governed by the product CR. The initial reverse output voltage is due to the forward coupling through the Miller capacitance but is small enough to be ignored. Then the transform of the output simplifies to:

$$E_5 = -\frac{I_A - I_B}{kCp(CRp+1)} \quad \ldots \quad \ldots \quad (9)$$

Ideally, the degenerative action of the Miller valve would keep the pick-off junction node (3) stationary at all times so that any slight bias on the diode electrodes [nodes (2) and (4)] would be sufficient to keep these valves shut off in the relaxation period and ensure that the two relaxation meshes remained independent. In practice, several volts of error signal have to be supported at node (3) in order to provide an appreciable drive to the Miller valve and to accumulate the steady bias that this valve requires. Fortunately, the charging impulses themselves produce relaxation waveforms on the diodes which serve to hold them off, provided the repetition period is not too long compared with the time constant CR. The minimum bias provided by the relaxation waveforms is quite simply estimated by considering the effect of a current impulse Q on a parallel CR combination. After a time τ the voltage across the capacity is given by:

$$e = \frac{Q}{C}\epsilon^{-\tau/CR} \quad \ldots \quad \ldots \quad (10)$$

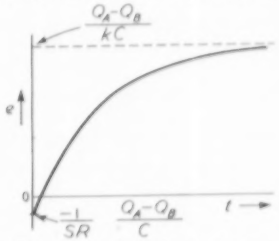


Fig. 23. *Impulse response of Fig. 22.*

If then the impulse is repeated again and again with uniform interval, the minimum voltage approaches the value

$$\frac{Q}{C}\frac{\epsilon^{-\tau/CR}}{1-\epsilon^{-\tau/CR}} \quad \ldots \quad \ldots \quad \ldots \quad (11)$$

If for example $\tau = CR$ then the minimum voltage in the steady state is $0{\cdot}581Q/C$ and, in the configuration of Fig. 20, this bias appears on the diode anodes and also, if $r \ll R$, on the cathodes. In the circuit of Fig. 13 the product CR is equal to the repetition period (500 μsecs) and, when signals are received through the i.f. amplifier, current impulses of the order of 5 mA and $\frac{1}{2}$ μsec duration are produced by V_{15}, which charge the 220-pF capacitors and increase the potential across them by about 10 V.

(*To be concluded*)

PARALLEL-T RC NETWORK

Simplified Analysis

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Introduction

THE parallel-T or twin-T resistance capacitance network in recent years has attained considerable popularity, mainly due to the fact that an attenuation characteristic similar to that of an *LC* network can be obtained without the use of an inductance. This advantage can be very important, particularly at very low frequencies. The network has been used for many years[1] and there is a comprehensive literature on its uses and properties[2,3,4,5,6]. There are many ways of analysing this network and one of the more recent papers, by L. G. Cowles[5], presents a most interesting and complete solution. However, it is felt that a more simple analysis which avoids the usual laborious mathematical solution would help design procedure and give some mental picture of its mode of operation which is essential when the effect of a variable element is to be considered.

The method adopted is to find an equivalent network of a more familiar pattern which is also more amenable to mathematical treatment. There are many ways of finding an equivalent network, one being the use of Bartlett's Bisection Theorem which leads to the familiar Wheatstone type of bridge.

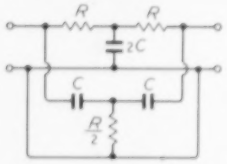

Fig. 1. *Basic parallel-T circuit; there is a null in the response when* $R = 1/\omega_0 C$.

1. Equivalent Network

The familiar parallel-T network is shown in Fig. 1, and the values of the components are chosen to give a null at a desired frequency (f_0) for the conditions shown. It is generally assumed that the network should work from a source of zero impedance into one of infinite impedance; these conditions simplify the mathematical analysis and give symmetrical characteristics. If the source and load impedances are finite, then the increased mathematical labours are seen from studying Fig. 2. Fig. 2(a) is fairly simply analysed by using 3-mesh equations but 4-mesh equations are necessary for the solution of Fig. 2(b). Having solved the equations, very little light is thrown on the principle of operation of the network.

However, an equivalent network can be found by making use of Bartlett's Bisection Theorem. The process, step by step, is shown in Fig. 3.

Step 1 shows Fig. 1.

Step 2 shows Fig. 1 bisected.

The Step 3 lattice network is obtained as follows:—The series arms of the lattice network consist of the input impedance of the network of Step 2 with all the bisected ends short-circuited; while the cross-arms of the lattice consist of the input impedance of Step 2 with all the bisected ends open-circuited. The process could readily stop at this point, as the lattice is a familiar network but Step 4 is more familiar and more easily understood.

Thus Step 4 is the final equivalent network which satisfies the conditions of simplified mathematical analysis and ease of understanding of its mode of operation.

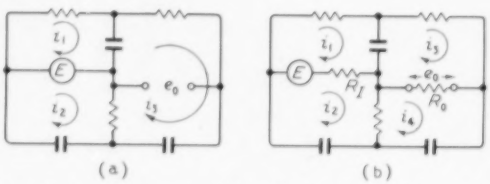

Fig. 2. *Mesh currents of parallel-T circuit when source is of zero impedance and load is infinite (a) and when both source and load are finite (b).*

2. Analysis

Examination of Fig. 3, Step 4 shows that if the network works from a zero-impedance source into one of infinite impedance, only two equations are necessary for its analysis. If the source and load impedance have finite values then three equations are necessary. This is seen from Fig. 4 where the impedances have been generalized. Although the source and load impedances have been shown as resistors they could also be of the form $R \pm jX$.

From Fig. 4 the mesh equations are:—

$$E = i_1 (R_I + Z_1 + Z_2) - i_2 Z_1 - i_3 Z_2 \quad \ldots \quad \ldots \quad (1)$$

$$0 = i_1 Z_1 + i_2 (R_0 + Z_1 + Z_2) - i_3 R_0 \quad \ldots \quad \ldots \quad (2)$$

$$0 = -i_1 Z_2 - i_2 R_0 + i_3 (R_0 + Z_1 + Z_2) \quad \ldots \quad \ldots \quad (3)$$

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From these equations the determinant is found which is:—

$$\Delta = (Z_1+Z_2)[2R_IR_0+(R_I+R_0)(Z_1+Z_2)+2Z_1Z_2] \quad \ldots \quad (4)$$

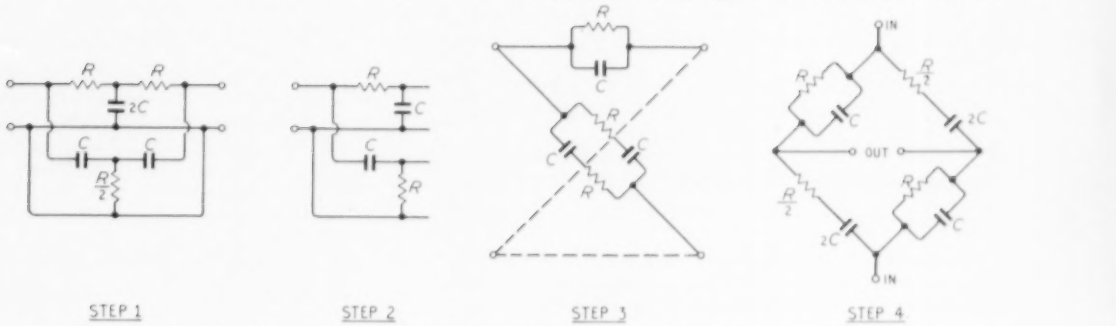

Fig. 3. *These diagrams illustrate the transformation of a parallel-T circuit through a lattice to a true bridge.*

The output voltage is

$$e_0 = (i_2 - i_3)R_0 \quad \ldots \quad \ldots \quad \ldots \quad (5)$$

where

$$i_2 = \frac{E[R_0Z_2 + Z_1(R_0 + Z_1 + Z_2)]}{\Delta} \quad \ldots \quad (6)$$

and

$$i_3 = \frac{E[R_0Z_1 + Z_2(R_0 + Z_1 + Z_2)]}{\Delta} \quad \ldots \quad (7)$$

Substituting (6) and (7) into (5) gives

$$e_0 = ER_0\frac{Z_1 - Z_2}{2R_IR_0 + (R_I+R_0)(Z_1+Z_2)+2Z_1Z_2} \quad (8)$$

Equation (8) completely defines the network, from which the attenuation and phase characteristics for any set of parameters can be plotted.

It is also very desirable to know the input impedance of the network.

Let Z_{in} = input impedance

From Fig. 4

$$Z_{in} = \frac{E}{i_1} - R_I \quad \ldots \quad \ldots \quad \ldots \quad (9)$$

where from equations (1), (2) and (3)

$$i_1 = \frac{E[(Z_1 + Z_2)(Z_1 + Z_2 + 2R_0)]}{\Delta} \quad \ldots \quad (10)$$

Therefore from (9) and (10)

$$Z_{in} = \frac{R_0(Z_1 + Z_2) + 2Z_1Z_2}{Z_1 + Z_2 + 2R_0} \quad \ldots \quad \ldots \quad (11)$$

3. Examination of Equation (8)

Rewrite equation (8) as:

$$\frac{E}{e_0} = \frac{2R_IR_0 + (R_I + R_0)(Z_1 + Z_2) + 2Z_1Z_2}{R_0(Z_1 - Z_2)} \quad \ldots \quad \ldots \quad (8a)$$

This equation can now be examined for various conditions.

3.1. Let $R_I = 0$ and $R_0 = \infty$. This is the normally-accepted working condition of this circuit.

Then $$\frac{E}{e_0} = \frac{Z_1 + Z_2}{Z_1 - Z_2} \quad \ldots \quad \ldots \quad \ldots \quad (12)$$

It is immediately apparent that a null will be obtained when $Z_1 = Z_2$. Some further properties will become apparent if the parameters of Fig. 3, Step 4 are substituted into equation (12).

Here $Z_1 = \dfrac{-jR}{\omega CR - j}$ and $Z_2 = \dfrac{\omega CR - j}{2\omega C}$

Therefore $$\frac{E}{e_0} = \frac{(\omega CR)^2 - 1 - j4\omega CR}{1 - (\omega CR)^2}$$

Let $R = 1/\omega_0 C$

Then $$\frac{E}{e_0} = \frac{(\omega/\omega_0)^2 - 1 - j4(\omega/\omega_0)}{1 - (\omega/\omega_0)^2} \quad \ldots \quad (13)$$

$$= \frac{\sqrt{\left[\{(\omega/\omega_0)^2 - 1\}^2 + \{4\omega/\omega_0\}^2\right]}}{1 - (\omega/\omega_0)^2}$$

with phase angle

$$-\tan^{-1}\frac{4\omega/\omega_0}{(\omega/\omega_0)^2 - 1} \quad \ldots \quad \ldots \quad (13a)$$

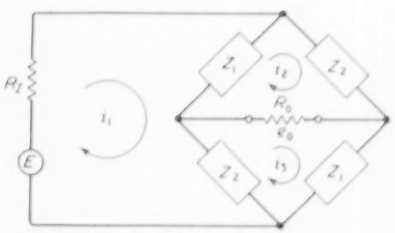

Fig. 4. *General bridge network.*

From (13a) the attenuation characteristic can be plotted against the parameter ω/ω_0, which is, in effect, against frequency (f) because f_0 is a constant. This method of presentation has many obvious advantages, one of the more important being compactness. From (13a) it can be seen that a null occurs when $\omega/\omega_0 = 1$ and also that the attenuation and phase characteristics are symmetrical about $\omega/\omega_0 = 1$. This last

point can be checked by substituting ω_0/ω for ω/ω_0 and equating the two equations; if they are identical, the characteristics are symmetrical.

3.2. Now let $R_I = R_I$ and $R_0 = \infty$

$$\text{Then } \frac{E}{e_0} = \frac{2R_I + Z_1 + Z_2}{Z_1 - Z_2} \quad \ldots \quad \ldots \quad (14)$$

Give values to Z_1 and Z_2 as in (3.1)

Then

$$\frac{E}{e_0} = \frac{(\omega CR)^2 - 1 + 4R_IR(\omega C)^2 - j(4\omega CR + 4\omega CR_I)}{1 - (\omega CR)^2}$$

Let $R = 1/\omega_0 C$

$$\frac{E}{e_0} = \frac{(\omega/\omega_0)^2(1 + 4R_I/R) - 1 - j4(\omega/\omega_0)(1 + R_I/R)}{1 - (\omega/\omega_0)^2}$$

$$\ldots \quad \ldots \quad (15)$$

Examination of (15) shows that a null occurs at $\omega/\omega_0 = 1$, but that the characteristics are not symmetrical. A rough shape of the attenuation characteristic can be found as follows:

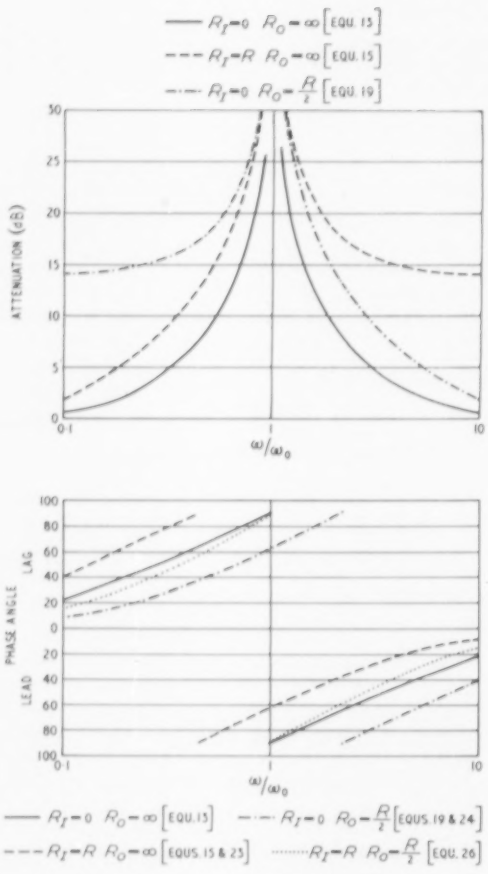

Fig. 5. *Normalized attenuation (a) and phase (b) characteristics of parallel-T network.*

$$\text{As } \frac{\omega}{\omega_0} \to \infty \quad \frac{E}{e_0} \to 1 + \frac{4R_I}{R} \quad \ldots \quad (16)$$

and as $\frac{\omega}{\omega_0} \to 0$

$$\frac{E}{e_0} \to -1 - j4\frac{\omega}{\omega_0}\left(1 + \frac{R_I}{R}\right) \to 1 \quad \ldots \quad (17)$$

Equation (17) compares with $(-1 - j4\omega/\omega_0)$ for the case when $R_I = 0$ (Section 3.1); both curves ultimately having the same value.

The attenuation and phase characteristics given by equations (13) and (15) are shown in Fig. 5(a) and (b) respectively. For equation (15) R_I has been given the value of R.

3.3. Let $R_I = 0$ and $R_0 = R_0$

$$\text{Then } \frac{E}{e_0} = \frac{R_0(Z_1 + Z_2) + 2Z_1Z_2}{R_0(Z_1 - Z_2)} \quad \ldots \quad (18)$$

$$\text{or } \frac{E}{e_0} = \frac{Z_1 + Z_2 + 2Z_1Z_2/R_0}{Z_1 - Z_2} \quad \ldots \quad (18a)$$

Assume the same values for Z_1 and Z_2 as used in (3.2).

The resulting expression is given by:

$$\frac{E}{e_0} = \frac{\{(\omega/\omega_0)^2 - 1 - 2R/R_0\} - j(\omega/\omega_0)(4 + 2R/R_0)}{1 - (\omega/\omega_0)^2}$$

$$\ldots \quad \ldots \quad (19)$$

Let us examine the general shape of the attenuation characteristic as follows:

At $f = \infty$, $Z_1 = 0$ and $Z_2 = R/2$

Therefore $E/e_0 = 1$ $\ldots \quad \ldots \quad \ldots \quad (20)$

At $f = 0$, $Z_1 = R$ and $Z_2 = \infty$

$$\text{Therefore } \frac{E}{e_0} = \frac{R_0 + 2R}{R_0} \quad \ldots \quad \ldots \quad \ldots \quad (21)$$

Compare equations (20) and (21) with (16) and (17) respectively. They appear to be inverse or mirror image curves and would be under certain conditions.

Equate (16) and (21):

$$1 + \frac{4R_I}{R} = \frac{R_0 + 2R}{R_0}$$

$$\text{Therefore } R = \sqrt{2R_IR_0} \quad \ldots \quad \ldots \quad \ldots \quad (22)$$

Now in (15) let $R_I = R$

$$\text{Therefore } \frac{E}{e_0} = \frac{5(\omega/\omega_0)^2 - 1 - j8\omega/\omega_0}{1 - (\omega/\omega_0)^2} \quad (23)$$

From (22) if $R_I = R$ then $R_0 = R/2$. Substitute this value into equation (19):

$$\frac{E}{e_0} = \frac{(\omega/\omega_0)^2 - 5 - j8\omega/\omega_0}{1 - (\omega/\omega_0)^2} \quad \ldots \quad \ldots \quad (24)$$

If expressions (23) and (24) give mirror image characteristics, then by substituting ω_0/ω for ω/ω_0 in (24), the resulting expression should be equivalent to (23). This has been done below and the equivalence shown. The differences in sign indicate phase lag or lead.

Then

$$\frac{(\omega/\omega_0)^2-5-j8\omega/\omega_0}{1-(\omega/\omega_0)^2} \rightarrow \frac{(\omega_0/\omega)^2-5-j8\omega_0/\omega}{1-(\omega_0/\omega)^2}$$

$$= \frac{1-5(\omega/\omega_0)^2-j8\omega/\omega_0}{(\omega/\omega_0)^2-1}$$

which shows that the expressions give mirror image characteristics. These results are shown in Fig. 5.

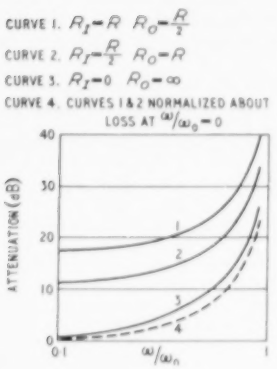

Fig. 6. *Attenuation characteristics symmetrical about $\omega/\omega_0=1$ for the condition $2R_IR_0=R^2$.*

It should now be possible to obtain a symmetrical characteristic with finite values of R_I and R_0.

3.4. Let $R_I = R_I$ and $R_0 = R_0$.
This expression was derived in Section 2 and is given by equation (8a) as

$$\frac{E}{e_0} = \frac{2R_IR_0 + (R_I + R_0)\ (Z_1 + Z_2) + 2Z_1Z_2}{R_0(Z_1 - Z_2)} \quad \ldots \quad \ldots \quad (8a)$$

Substitute values of Z_1 and Z_2 as in Section 3.1:.
Therefore

$$\frac{E}{e_0} = \left[\left(\frac{\omega}{\omega_0}\right)^2\left(\frac{4R_I}{R}+\frac{R_I}{R_0}+1\right)-\left(\frac{R_I}{R_0}+\frac{2R}{R_0}+1\right)\right.$$
$$\left.-j\frac{\omega}{\omega_0}\left\{\frac{4R_I}{R}+4\left(\frac{R_I}{R_0}+1\right)+\frac{2R}{R_0}\right\}\right]\Big/\left[1-\left(\frac{\omega}{\omega_0}\right)^2\right] \quad \ldots \quad (25)$$

For this curve to be symmetrical the condition of equation (22) must hold.
Substitute $R = \sqrt{2R_IR_0}$ into (25).
Then

$$\frac{E}{e_0} = \left[\left\{\left(\frac{\omega}{\omega_0}\right)^2-1\right\}\left\{\frac{R_I}{R_0}+2\sqrt{\left(\frac{2R_I}{R_0}\right)}+1\right\}\right.$$
$$\left.-j4\left\{\frac{R_I}{R_0}+\sqrt{\left(\frac{2R_I}{R_0}\right)}+1\right\}\frac{\omega}{\omega_0}\right]\Big/\left[1-(\omega/\omega_0)^2\right] \quad \ldots \quad (26)$$

It is apparent that this expression will give a symmetrical characteristic as it is of a similar form to equation (13). The basic loss will always be greater than that of equation (13), the degree depending on the values of R_I and R_0, but when normalized it will be seen that a sharper characteristic is obtained. This is shown in Fig. 6 where curve 3 shows the response for $R_I = 0, R_0 = \infty$; curve 2 for $R_I = R/2, R_0 = R$; curve 1 for $R_I = R, R_0 = R/2$ and curve 4 is the normalized result of curves 1 and 2. It is seen that the conditions chosen for curves 1 and 2 give normalized curves that are coincident. If equation (26) is normalized, then

$$\frac{E}{e_0} = \frac{(\omega/\omega_0)^2 - 1 - j4(\omega/\omega_0)(A/B)}{1-(\omega/\omega_0)^2} \quad \ldots \quad (27)$$

where $A = \frac{R_I}{R_0} + \sqrt{\left(\frac{2R_I}{R_0}\right)} + 1$

and $B = \frac{R_I}{R_0} + 2\sqrt{\left(\frac{2R_I}{R_0}\right)} + 1$

and it can be shown that the ratio A/B has a minimum value when $R_I/R_0 = 1$ and further is symmetrical about this point. Therefore, the sharpest characteristic will occur at $R_I/R_0 = 1$ and this is the reason for the coincidence of curves 1 and 2 when normalized.

4. General Circuit

The foregoing analysis has been based on Fig. 3 where, as shown in Step 2, the components in the two T-arms have similar values. A more general circuit would be of the form shown in Fig. 7 where "n" is any number. The null condition is given by $Z_1 = Z_2$ and, if the values of Fig. 7 are used, it is found that a null occurs when

$$R = 1/\omega_0 C\sqrt{n} \quad \ldots \quad \ldots \quad \ldots \quad (28)$$

By inserting the values of Z_1 and Z_2 into the appropriate formulae given in previous sections any desired result can be obtained. The main properties of this circuit can be examined if it is compared with the case of $n = 1$, $R_I = 0$, $R_0 = \infty$ (Section 3.1).

By introducing the values of Z_1 and Z_2 of Fig. 7 into equation (12) the following result is obtained:

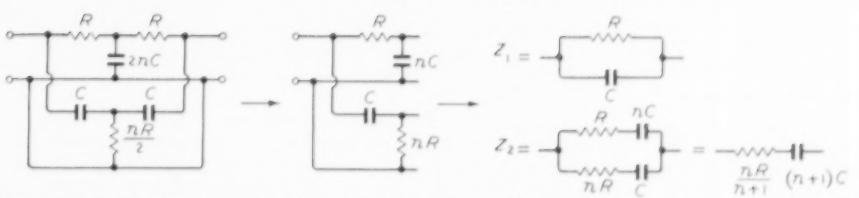

Fig. 7. *Derivation of bridge arms when the shunt arms of the parallel-T circuit are n times as great as in Fig. 1.*

$$\frac{E}{e_0} = \frac{\left(\frac{\omega}{\omega_0}\right)^2 - 1 - j\frac{2}{\sqrt{n}}(n+1)\frac{\omega}{\omega_0}}{1 - (\omega/\omega_0)^2} \quad \ldots \quad (29)$$

By examining the constant of the j term it is seen that it is a minimum when $n = 1$ and that values are symmetrical about $n = 1$ (that is, at $n = 0{\cdot}25$ and $4{\cdot}0$, etc.), so that the sharpest attenuation characteristic is obtained when $n = 1$.

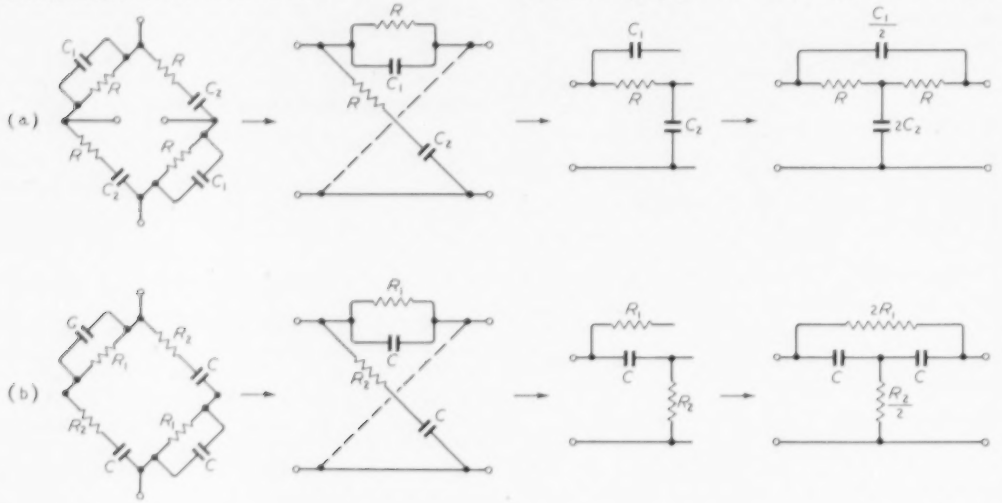

Fig. 8 (above). *This diagram illustrates how the bridge form of the circuit can be transformed to the bridged-T.*

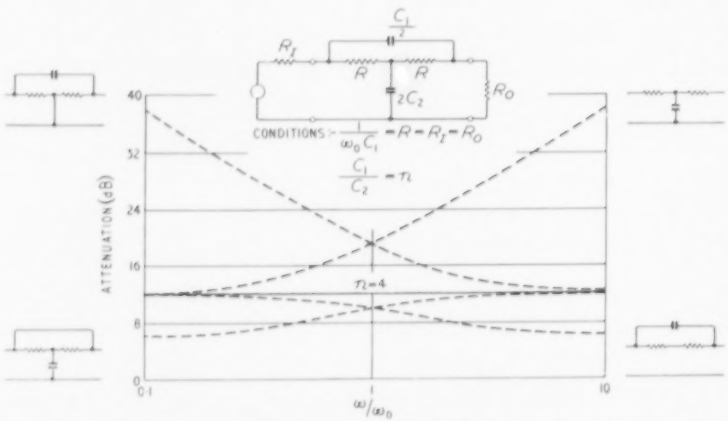

Fig. 9 (right). *Amplitude characteristics of Fig. 8 (a) for* $R = R_0 = R_I = 1/\omega_0 C_1$.

When R_I and R_0 have finite values, an expression of a similar form to equation (25) can be obtained and, for the attenuation characteristic to be symmetrical, it is found that

$$R_I R_0 = R^2 \frac{n}{n+1} \quad \ldots \quad \ldots \quad \ldots \quad (30)$$

5. Circuit Variations

From a study of Fig. 3 it is apparent that starting with any other configuration of Step 4, the procedure can be reversed, ending at Step 1 with a circuit that may be appreciably different from that of the normal parallel-T. Such a procedure is shown in Fig. 8; the resulting circuits being of the bridged-T variety. These circuits have some very useful properties which can be easily examined by introducing the appropriate values of Z_1 and Z_2 into equation (8). An example of this is shown in Fig. 9 which shows the variations of amplitude characteristic obtainable for the circuit of Fig. 8(a) for the condition $1/\omega_0 C_1 = R = R_0 = R_I$. If in Fig. 9, $n = 4$, a characteristic is obtained independent of attenuation; so that this circuit could be used for eliminating the effects of an unwanted capacitance on an attenuation characteristic.

6. Conclusions

It has been shown that the design of a parallel-T network can be simplified considerably by this method of equivalence. A desired characteristic can be obtained by either varying the

source and load impedance or by varying the ratio of the parallel-T components (that is, the value of n). Finally, it is seen that variations of the parallel-T or bridged-T can be simply analysed and hence designed precisely.

Acknowledgment

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D.F. PLOTTING AID

By H. G. Hopkins, B.Sc., Ph.D., M.I.E.E.

(Official Communication from D.S.I.R. Radio Research Station, Slough)

SUMMARY.—A simple plotting aid is described for use in determining the contour enclosing the area within which the transmitter lies with a given probability. It is in the form of a transparent graticule which is placed on the map over the most probable point and is designed to facilitate the calculations. The aid is flexible in application and suitable for use where speed is required rather than the highest accuracy.

1. Introduction

WHEN a network of direction finders is used to determine the position of a distant transmitter, each instrument provides a best estimate of bearing, due allowance having been made for known systematic errors. The residual random error associated with the bearing provided by the jth direction finder may be represented by the standard deviation σ_j; the standard deviations can be separately estimated. Stansfield[1] has shown how to determine the most probable position (m.p.p.) of the transmitter from these data and that, granted certain assumptions of a geometrical nature, the contour enclosing the *minimum* area on the earth's surface within which the transmitter lies with a given probability is elliptical in shape and centred on the m.p.p. His analysis enables the direction and magnitudes of the ellipse axes to be computed. The computation is rather lengthy and Barfield[2] has suggested a plotting procedure which is speedier and avoids some of the complications involved; the aid described in the present note has a similar objective. A balance of the conflicting factors of precision and time will determine which method of determining the probability contours is best for any particular purpose. Limited experience of the technique to be described suggests that it is quicker than that proposed by Barfield, at any rate for those not fully experienced in his method. The accuracies of the two methods appear to be comparable.

2. Basis of the Method

It is assumed that the most probable position, F, of the transmitter has already been determined and marked on a gnomonic plotting chart showing the positions of the direction finders. Let orthogonal axes be set up at F; to fix ideas let the Y axis be along the meridian at F and let angles be measured clockwise from north. Suppose θj to be the angle between the meridian at F and the bearing-string from the jth direction finder when the string is positioned over F. Following Stansfield let

$$\lambda = \sum \frac{\sin^2\theta_j}{d_j^2\sigma_j^2},$$

$$\mu = \sum \frac{\cos^2\theta_j}{d_j^2\sigma_j^2}, \text{ and}$$

$$\nu = \sum \frac{\sin\theta_j \cos\theta_j}{d_j^2\sigma_j^2}$$

where d_j is the length of the string between the jth direction finder and F, and where the summations are carried out over all the direction finders. Then the constants for the ellipse

$$\frac{x^2}{b_p^2} + \frac{y^2}{a_p^2} = 1$$

centred on F and enclosing the area on the chart associated with a given probability, p, are approximately given by:

$$\tan 2\phi = -\frac{2\nu}{\lambda - \mu} \quad \ldots \quad \ldots \quad \ldots \quad (1)$$

and, taking the root $-\frac{\pi}{4} < \phi < \frac{\pi}{4}$,

$$-\frac{2\log_e(1-p)}{a_p^2} = \lambda - \nu\tan\phi \quad \ldots \quad (2)$$

and $$-\frac{2\log_e(1-p)}{b_p^2} = \mu + \nu\tan\phi \quad \ldots \quad (3)$$

MS accepted by the Editor, July 1955

In these equations a_p and b_p are the (orthogonal) semi-axes of the ellipse measured in the same units as d_j, and ϕ is the angle between the a_p axis and the meridian at F. Note that a_p is not necessarily the major semi-axis.

These equations correspond to (12), (15) and (16) of Stansfield, corrected in the case of (15) and (16) to allow for erroneous factors of 2 in the λ and μ terms which appear to have arisen in the step following equation (14).

The difficulty in quickly applying the above equations lies in the determination of λ, μ and ν. This may be facilitated as follows.

Let the string centred on the jth direction finder be rotated clockwise by an angle σ_j from the position it occupies when aligned on F, the angle σ_j being as indicated on the bearing rose for that direction finder. If y_j is the intercept of the deflected string on the Y axis, it will be noted that y_j^{-2} is approximately equal to the contribution to λ of the jth direction finder, namely, $\dfrac{\sin^2 \theta_j}{d_j^2\, \sigma_j^2}$; similarly the contribution to μ is x_j^{-2} where x_j is the intercept on the X axis.

Thus, if the process described above is carried out for each string in turn and the x_j and y_j noted, then λ and μ can be derived by summation. The jth contribution to ν is given by $x_j^{-1} y_j^{-1}$ and, taking due note of signs, ν itself is found by summation.*

These relationships are applied in the construction of the simple plotting aid described below.

3. Description of the Plotting Aid

The plotting aid is shown in Fig. 1. It comprises orthogonal axes each graduated so as to read directly the reciprocal of the square of the distance from the origin. The graticule is engraved on the upper surface of a flat, rigid sheet of polished transparent material having chamfered edges to reduce friction with the strings. The region near the intersection of the axes is of little value because the scale is very crowded; the central portion of the upper surface is accordingly left blank, but a cross is engraved, as shown immediately below the centre of the graticule. This cross enables the aid to be set over the m.p.p. without parallax. The positive and negative signs in the four quadrants are used, as described later.

* Mr. E. M. L. Beale has kindly indicated to the author that ν can also be found by using an auxiliary set of orthogonal axes (X_a, Y_a) rotated clockwise by $\pi/4$ from the (X, Y) set. If x_{aj} and y_{aj} are the intercepts of the deflected jth string on X_a and Y_a, then $2\nu = \Sigma(x_{aj}^{-2}) - \Sigma(y_{aj}^{-2})$. The relation $\Sigma(x_j^{-2}) + \Sigma(y_j^{-2}) = \Sigma(x_{aj}^{-2}) + \Sigma(y_{aj}^{-2})$ also provides a useful computational check.

A convenient size for the transparent sheet has been found to be about 25 cm square; this enables the scales along the axes to cover the range 0·01 to 2 cm^{-2} where the centimetre is used as the unit for measuring the axial dimensions of the ellipses. The unit of length used in Fig. 1 has been chosen arbitrarily. To incorporate Mr. Beale's suggestion (mentioned in the footnote) would involve engraving on the sheet the second set of axes, similarly graduated. Since no experience has been gained of the aid so modified, it is not considered further.

In use, the aid is centred on the m.p.p. (assumed to be already known) and adjusted so that the s^2 (or Y axis) is along the meridian at this point. The choice of the meridian as reference direction is mainly for convenience of description in this account; in practice, it may sometimes be found desirable to orient the graticule differently, for example, as a means of improving reading accuracy. Each string is taken in turn and aligned on the m.p.p.; the

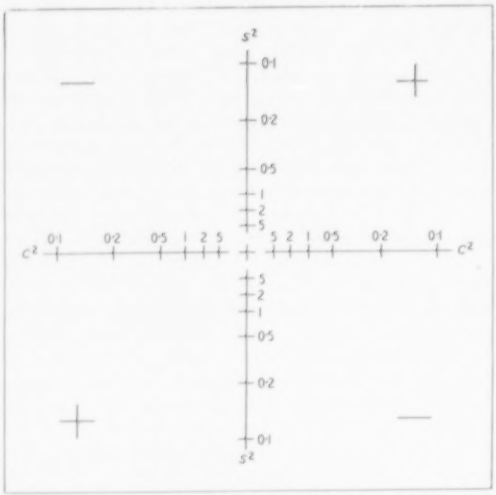


Fig. 1. *Diagram of plotting aid; a few graduations are shown to illustrate the scale.*

bearing of the m.p.p. is read off on the appropriate bearing rose and the string is rotated clockwise by its σ_j. The co-ordinates l_j, m_j of the points where the string crosses the s^2 and c^2 axes are read off from the scale graduations and the sign of the quadrant across which the string cuts between the intercepts is noted. The results are set out in the s^2, c^2 and Sign columns as shown in the Table.

The entries in the s^2 and c^2 columns are all essentially positive; the scale readings are directly entered as positive values independently of how the deflected strings cut the axes. The

sign read off as mentioned above applies to the entries to be made in the *sc* column; these may be conveniently read off from a nomogram giving the geometric mean of l_j and m_j or may be calculated from these values. The former process is quicker and avoids mistakes involving $\sqrt{10}$ which can easily occur in the latter.

The constants of the ellipse enclosing the area associated with probability p are then given by:

$$\tan 2\phi = \frac{2N}{L - M} \quad \ldots \quad \ldots \quad \ldots \quad (4)$$

and, taking the root $-\frac{\pi}{4} < \phi < \frac{\pi}{4}$

$$-\frac{2 \log_e (1 - p)}{a^2_p} = L + N \tan \phi \ldots \quad (5)$$

and $$-\frac{2 \log_e (1 - p)}{b^2_p} = M - N \tan \phi \quad (6)$$

It will be recalled that ϕ (the direction of the a_p axis) is measured clockwise from the s^2 axis and that a_p and b_p are in the units used in constructing the graticule: a_p is not necessarily greater than b_p. The discrepancies in sign between these three working equations and (1), (2) and (3) are a result of the signs allotted to the quadrants of the plotting aid. The scheme adopted (Fig. 1) appears more natural than the alternative, implicit in the earlier equations, in which the quadrants would have the opposite signs. It must be stressed that equations (4), (5) and (6) must be used if the quadrants are labelled as shown in the figure.

With but little practice any chosen probability contour can be determined in less than ten minutes when four direction finders are in use. Limited experience suggests that the directions of the axes are usually accurate to one or two degrees and the axial lengths to within about 20% except when long thin ellipses are involved. In this case the errors in the long axis may be very large but may be somewhat reduced by orienting the s^2 axis roughly along the expected direction of the long axis rather than along the meridian. A useful practical way of improving accuracy when the intercepts on the axes are small is to multiply *all* σ_j by a convenient factor k (exceeding unity) thus removing the deflected strings from the crowded region of the scale near the origin; ϕ is unaffected but the derived values of a_p and b_p should be divided by k.

TABLE

Direction Finder	Scale Reading s^2	Scale Reading c^2	Sign	Derived *sc*
1	l_1	m_1		$\sqrt{(l_1 m_1)}$
2	l_2	m_2		$\sqrt{(l_2 m_2)}$
3	l_3	m_3		$\sqrt{(l_3 m_3)}$
...	...	...		...
...	...	...		...
	Sum = L	Sum = M		Sum = N (algebraic)

4. Conclusion

The aid described is put forward as a practical and simple means of rapidly, if roughly, determining the elliptical area about the most probable point within which the transmitter lies with a chosen probability. It is useful on a map of any scale since the axial lengths are in units determined by the graduations on the aid itself and not by the scale of the map.

It is opportune to consider simple plotting aids of this general character since the rather complicated and lengthy computations inherent in statistical plotting techniques appear to have constituted one of the stumbling blocks in the way of the widespread adoption of these objective methods.

Acknowledgment

The work described above was carried out as part of the programme of the Radio Research Board. This paper is published by permission of the Director of Radio Research of the Department of Scientific and Industrial Research.

REFERENCES

1 R. G. Stansfield, "Statistical Theory of D.F. Fixing", *J. Instn elect. Engrs*, Part IIIA, 1947, Vol. 94, p. 762.

2 R. H. Barfield, "Statistical Plotting Methods for Radio Direction Finding", loc. cit., p. 673.

CORRESPONDENCE

Letters to the Editor on technical subjects are always welcome. In publishing such communications the Editors do not necessarily endorse any technical or general statements which they may contain.

New Rejector Circuit

SIR,—The April Editorial sets out to explain the behaviour of the so-called "bifilar-T rejector" by deriving an equivalent ladder circuit, free from mutual inductance. The equivalent, which appears in Fig. 4 of the Editorial, contains a tuned circuit with negative components, and it is to this feature that the special properties of the circuit are ascribed. The alternative treatment given below shows that these negative components are simply an incidental characteristic of the particular equivalent chosen and have no connection with the essential properties of the rejector.

The left-hand side of Fig. 1 shows the rejector circuit in question with the trivial modification of having a resistor of $4R$ placed across the whole of the tightly-coupled coil rather than a resistor of R across one half it; this modification is perfectly legitimate in view of the assumed perfect coupling and it probably represents a better arrangement for the practical case of slightly imperfect coupling. On the right-hand side of Fig. 1 appears the lattice network which is equivalent to the rejector. This lattice equivalent may be found from purely topological considerations using the methods described by Bode[1].

If we postulate $R = R_1$ then resistances of $2R_1$ may be removed in shunt from both lattice arms and placed in parallel with the input and output terminals as in Fig. 2. At this stage we can see the true significance of the bifilar rejector—it allows us to remove all the dissipation

from the network and place it across the source and load impedances. What is left is a network of dissipationless reactances which can consequently exhibit sharp discrimination and have true infinite-loss points at real frequencies. This technique has been used in conventional filters for many years and has been described by Mason[2].

networks. Although the circuit is potentially non-minimum-phase, as used it has minimum-phase properties as Fig. 3 indicates, and the phase characteristic is precisely the same as that of any other minimum-phase network with the same loss characteristic.

The problem of providing a band-pass response with

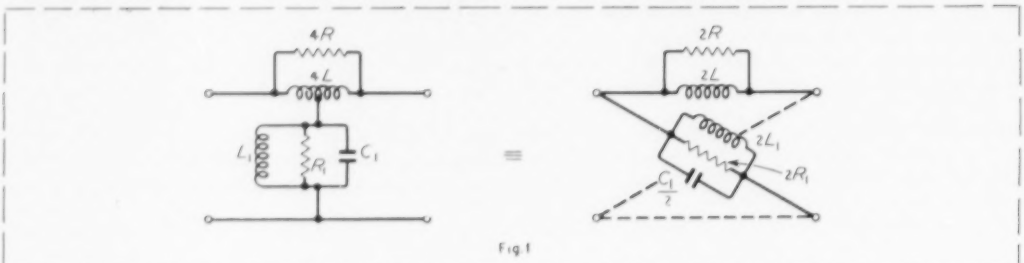


Fig. 1

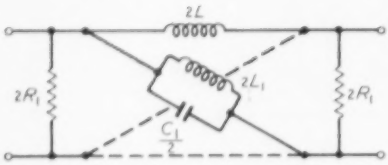


Fig. 2

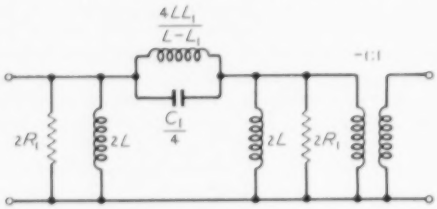


Fig. 3

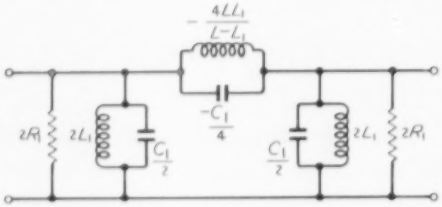


Fig. 4

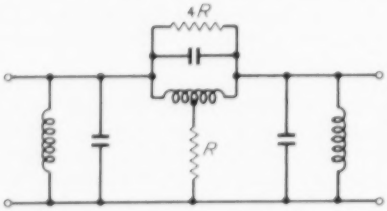


Fig. 5

In this particular case, the lattice reactance network is easily recognized as a simple high-pass filter of which the ladder equivalent is given in Fig. 3. The phase inversion which the original circuit provides at very high frequencies is produced in the equivalent by means of the tandem-connected ideal transformer. If, for some reason, one wishes to avoid the presence of transformers in *any* form in the unbalanced equivalent it is of course possible, but one is then forced to use negative elements in order to obtain the phase inversion. Such an equivalent is shown in Fig. 4, and is the circuit derived in the Editorial. By comparison with Fig. 3 it is more complicated and it is almost impossible to see clearly how the circuit works. Evidently the negative components have nothing to do with the sharp discrimination which the rejector may give.

There are many other circuits which can have their dissipation effects removed in the same way as the "bifilar-T". A band-pass example is given in Fig. 5. In the article[3] where the "bifilar-T" circuit was described by Fairhurst it was stated to be a non-minimum-phase network and this fact was said to be responsible for its superior phase characteristic compared with those of other equally-selective but otherwise minimum-phase sharp rejection near one edge of the pass-band, as needed for a vision i.f. channel, has been solved by conventional filter theory many years ago. Designing a simple band-pass filter so that its effective pass-band coincides with the wanted vision channel and with a peak of loss at the sound-channel frequency gives a circuit as shown in Fig. 5 in which all the dissipation has been thrown across the terminating resistors by using the resistance-compensation trick described by Mason. This circuit has the advantage of giving capacitances across input and output terminals, which can be used to absorb the valve capacitances, and also across the bifilar-wound inductor to absorb its self-capacitance.

H. J. Orchard

Post Office Research Station,
Dollis Hill,
London, N.W.2.
21st May 1956.

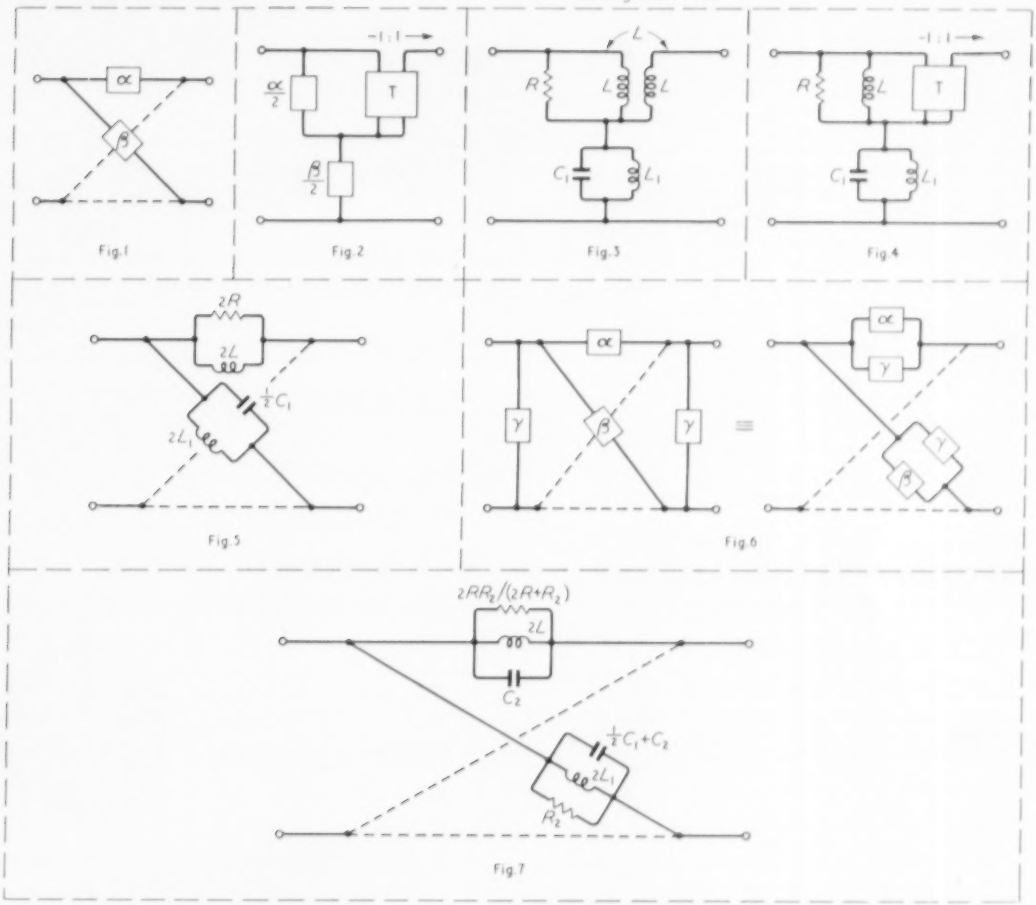
REFERENCES

[1] H. W. Bode, "Network Analysis and Feedback Amplifier Design", pp. 266–270, D. Van Nostrand, New York, 1945.
[2] W. P. Mason, "Resistance Compensated Band-Pass Crystal Filters for use in Unbalanced Circuits", *Bell Syst. Tech. J.*, Vol. 16, p. 423, 1937.
[3] H. A. Fairhurst, "Experimental Color Receiver", *Wireless World*, March 1956, p. 112.

Bifilar-T Trap

SIR,—I have been very interested to read your leaders in *Wireless Engineer* (April and May) on the "bifilar-T trap". I should like, however, to suggest to you that the equivalent network of the device can be obtained with great simplicity by recourse to two known theorems due respectively to Cauer and Bode. Both of these are most readily derived by matrix theory.

Cauer showed, in about 1930, that the lattice network, Fig. 1, is equivalent to the network of Fig. 2, in which T is an ideal transformer.



Now the bifilar-T trap, as shown in your Fig. 1(a), p. 105, May, and here re-drawn in the form of Fig. 3, is obviously equivalent to Fig. 4 when T again is an ideal transformer of ratio $-1 : 1$.

It follows that this network is then equivalent at all frequencies to the lattice (bridge) network of Fig. 5.

Coming now to the effect of bilateral shunting of the network by leaking capacitors, we can make use of the equivalence (shown by Bode* in 1934) illustrated in Fig. 6. From this it follows immediately that the bifilar-T trap, terminated as in your Fig. 3, p. 106, with, say, C_2 in parallel with R_2, is identically equivalent to the lattice network of Fig. 7, for which the conditions of maximum (or infinite) attenuation and overall transmission band can be calculated from the usual lattice-network relations.

The above lattice equivalents, it is interesting to note, have no "negative elements" and are therefore realizable physically.

W. PROCTOR WILSON

The British Broadcasting Corporation,
Kingswood Warren, Surrey.
14th June 1956.

* "General Theory of Electric Wave Filters", *Journal of Mathematics and Physics*, Vol. 13, No. 3, November 1934, p. 341.

INTERNATIONAL SCIENTIFIC RADIO UNION

The Proceedings of the 11th General Assembly of U.R.S.I. which was held on 23rd August–2nd September 1954 can be obtained from the General Secretariat of the U.R.S.I., 42 Rue des Minimes, Brussels, Belgium, as follows: Vol. 10, Part 2, Commission 2 on Radio and Troposphere, price 14s. 6d.; Part 7, Commission 7 on Radioelectronics, price 31s. 6d.

"Phase-Angle Measurement"

The author, C. H. Vincent, of this article in the May issue, is with the Engineering Department of the University of Edinburgh.

MSF STANDARD FREQUENCIES EXPRESSED IN TERMS OF THE CAESIUM RESONANCE

By L. Essen, D.Sc., Ph.D., A.M.I.E.E.

(Communication from the National Physical Laboratory)

As reported earlier[1] the MSF transmissions have been measured in terms of the caesium resonator since June 1955. No values have yet been published because the caesium frequency had been determined[2] only in terms of the unit of provisional uniform time for June of that year. It has been found[3], however, that this value differs very little from the value expressed in the unit of provisional uniform time averaged over the year June 1955–June 1956; and it has therefore been decided at the National Physical Laboratory to use this value 9 192 631 830 c/s as a provisional standard of frequency. Expressed in terms of this unit the mid-monthly deviations of the MSF 60-kc/s transmissions received at Teddington are as follows:—

Date	Deviation parts in 10^9	Date	Deviation parts in 10^9
*June 1955	−5	Dec. 1955	+1
July	−2	Jan. 1956	−1
Aug.	−1	Feb.	−1
Sept.	0	March	+1
Oct.	+1	April	+1
Nov.	+1	May	+3

* Value for 20th June. Small variation on 18th June.

In order to help those users who require the highest precision, for the testing of the drift rates of quartz standards for example, the values of MSF published in *Wireless Engineer* will, in future, be expressed in terms of this atomic standard of frequency with a precision of ± 1 part in 10^9. The transmission frequency itself will be controlled to within ± 5 parts in 10^9 of this frequency, being adjusted as before by steps of probably ± 5 parts in 10^9 when necessary. (No adjustment has been made since 6th May 1953.)

This first practical application of an atomic frequency standard is particularly significant, because other atomic standards are likely to come into operation soon and these transmissions, with the help of the 16-kc/s GBR transmission which is controlled by the same quartz oscillator, will enable them to be measured directly in terms of the N.P.L. caesium resonator, wherever they are situated.[4]

The speech announcement made during the transmissions has been modified and is now as follows:—

> This is MSF Rugby transmitting carrier and modulation frequencies within 5 parts in 10^9 of the nominal values on 2 500, 5 000 and 10 000 kc/s, as determined by an atomic standard. A transmission of one hour duration is also radiated on 60 kc/s at 14·29 GMT. Carriers are modulated with 1 000 c/s for the first five minutes and with 1 c/s pulses for the second five minutes of each quarter hour. At 15 minutes past each hour all transmissions are interrupted for five minutes. Enquiries should be directed to the National Physical Laboratory, Teddington, Middlesex, England. This is MSF, Rugby, England.

Reception reports are no longer requested because it is considered that the coverage of the service is now well established. The reports sent in the past have been most useful for this purpose. Any suggestions or comments about the service are, however, still welcomed and should be addressed to the Director, National Physical Laboratory, Teddington, Middlesex, England.

This note is published by permission of the Director of the National Physical Laboratory.

REFERENCES

1 L. Essen, *Wireless Engineer*, Vol. 32, p. 312, 1955.
2 L. Essen and J. V. L. Parry, *Nature*, Vol. 176, p. 280, 1955.
3 L. Essen and J. V. L. Parry, *Nature*, Vol. 177, p. 744, 1956.
4 J. A. Pierce, H. T. Mitchell and L. Essen, *Nature*, Vol. 174, p. 922, 1954.

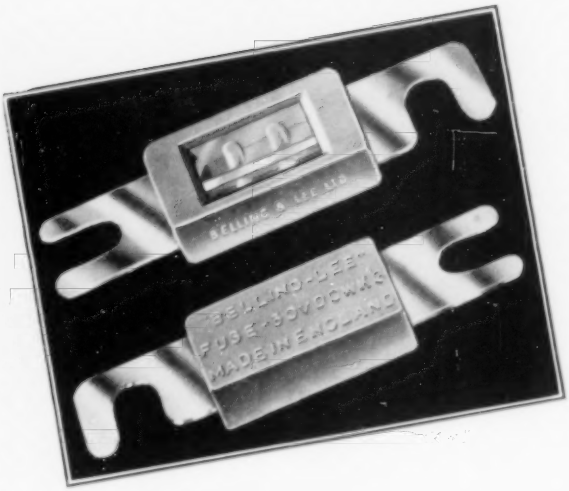
STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Values for May 1956

Date 1956 May	Frequency deviation from nominal: parts in 10^8	
	MSF 60 kc/s 1429-1530 G.M.T.	Droitwich 200 kc/s 1030 G.M.T.
1	+0·4	−1
2	+0·5	+4
3	+0·5	−1
4	+0·5	+3
5	N.M.	−1
6	N.M.	N.M.
7	+0·5	−2
8	+0·4	−1
9	+0·5	−1
10	+0·4	0
11	+0·5	−1
12	+0·5	−1
13	+0·5	0
14	+0·5	+3
15	+0·5	−1
16	N.T.	+4
17	N.T.	−1
18	+0·5	+5
19	+0·5	N.M.
20	+0·5	N.M.
21	+0·6	−3
22	+0·6	−1
23	+0·6	−2
24	+0·6	−1
25	+0·6	−1
26	+0·6	−1
27	+0·6	−2
28	+0·5	−1
29	+0·5	−2
30	+0·5	−1
31	+0·6	−2

The values are based on astronomical data available on 1st June 1956.
The transmitter employed for the 60-kc/s signal is sometimes required for another service.
N.M. = Not Measured N.T. = No Transmission



AIRFUSE
L.1330

CONSTRUCTION:

The fusible element is of electro-tinned copper and is in one piece with end connections which are suitably shaped for direct bolting to bus-bars, etc. The body is of high grade ceramic and is fitted with a glass window so that the fusible part of the element is visible. The rating is stamped on the lugs and is also printed on the glass.

SPECIFICATION:

The fuselink is to be capable of breaking currents of up to 3,000 amperes at 30V d.c. (2,500 amperes at 30V. d.c. for the 35A and 50A ratings) without shattering. The arcing time shall not exceed 0.003 seconds at 3,000 amperes.

The fuse is intended as a direct replacement for the American "Current Limiter" used on 28V. d.c. aircraft power supplies, and for this purpose has received the approval of the Air Registration Board under their reference No. E.3217.

It is also suitable for the protection of battery operated vehicles, heavy current rectifier output circuits, low voltage furnaces and other similar d.c. applications.

RATINGS:

35A, 50A, 80A, 100A, 130A, 150A, 200A, 225A, and 275A. The range will be extended in four further steps to 500A maximum.

HEAVY CURRENT, LOW VOLTAGE, FUSE-LINKS

for use on D.C. supplies in
AIRCRAFT, RECTIFIER EQUIPMENT, BATTERY DRIVEN VEHICLES, BATTERY CHARGING PLANT, &c.

H.R.C. CERAMIC BODIED FUSE-LINKS

The high rupturing capacity fuse-links comply with the requirements of R.C.S. 261 and B.S. 88 — made in two sizes, and a range of seven ratings in each size, as follows:—

L.693. 1½" x ¼". Ratings: 0.25, 0.5, 1, 2, 3, 5 & 7A. Category of duty: B.S. 88, a.c.4. 440V. a.c. at 55°C. It is expected that their d.c. category of duty will comply with B.S. 88, d.c.4. 230V.

L.754. ¼" x 3 1/16". Ratings: 0.025, 0.050, 0.100, 0.250, 0.500, 1 & 2A. Category of duty: B.S. 88, a.c.2. & d.c.2., 250V a.c. and 230V. d.c. at 55°C.

Also available but not to the above specifications:

L.760. 1½" x ¼" Ceramic, unfilled, designed to blow within 10 secs. on 100% overload.

BELLING & LEE LTD
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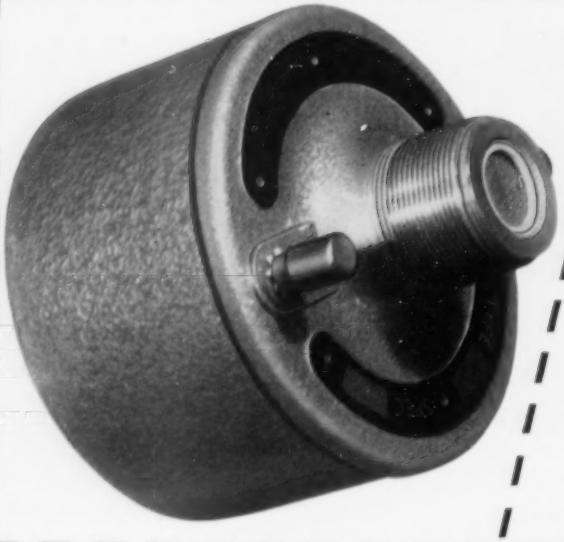

DC 10
PRESSURE UNIT

SPECIFICATION

Power Handling Capacity	10 watts peak
Voice Coil Impedance	15 ohms
Frequency Response	120–9000 c.p.s.
Flux Density	12,000 gauss
Pole Piece	1·5" diameter

Price £6 10s. 0d.

DIMENSIONS

	DC 10 (without line trs.)	DC 12 (with line trs.)
Diameter	4⅛"	4⅛"
Length	4¼"	6⅝"
Weight	4¾lb.	5½lb.

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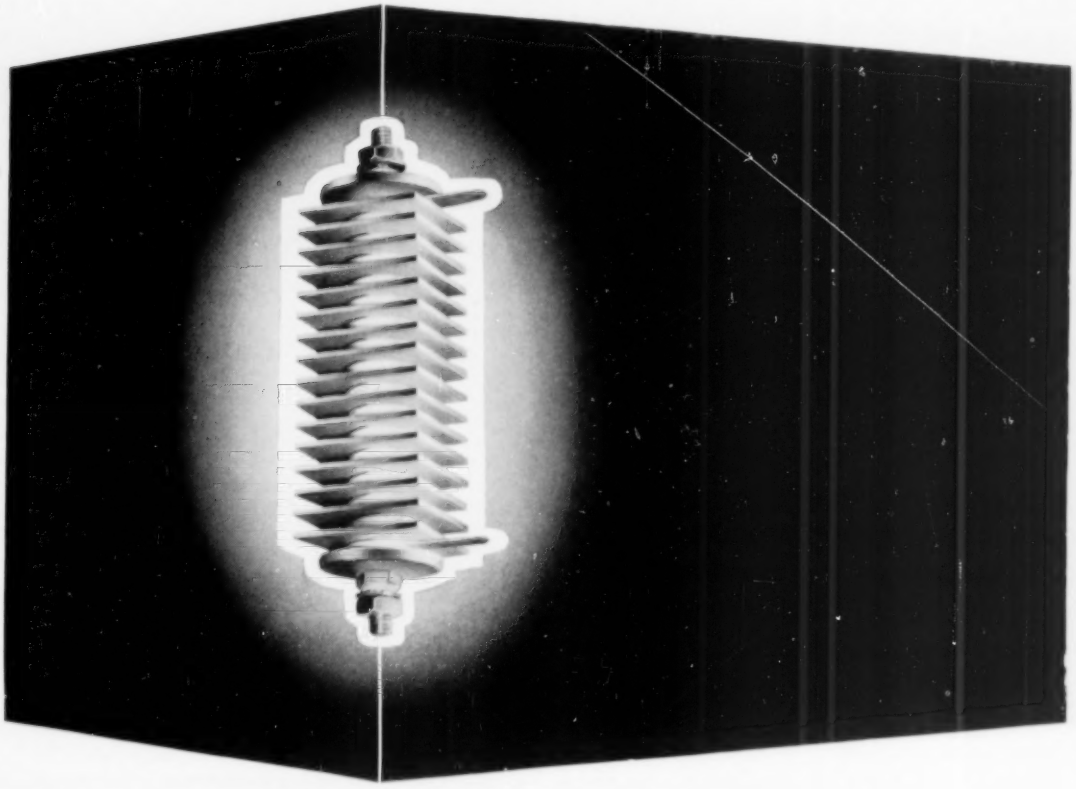

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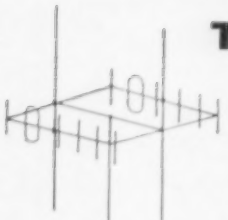

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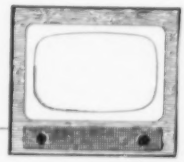

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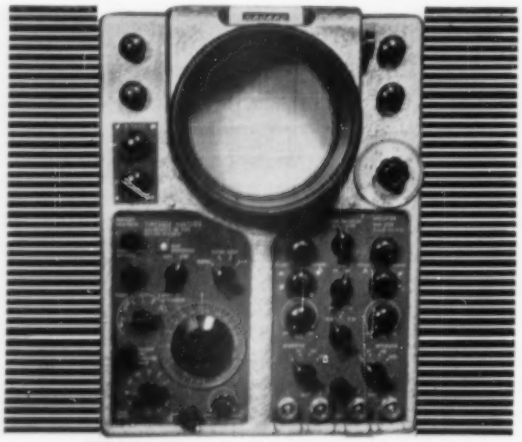

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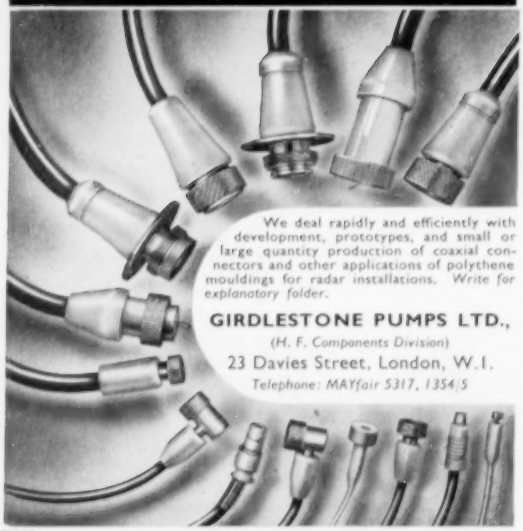

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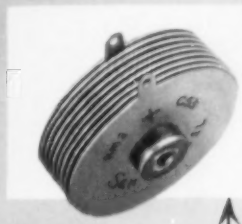

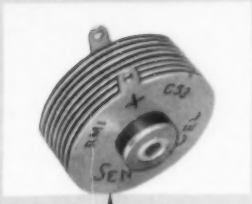

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Maximum output current (mean)	30mA 15mA	60mA 30mA	100mA 60mA	120mA 90mA	250mA 125mA	300mA 150mA
Maximum input voltage (r.m.s.)	125V	125V	125V	125V	250V	250V
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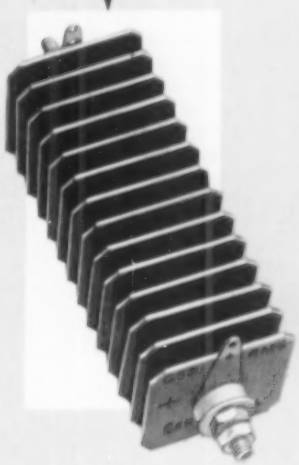
*For use in voltage doubler circuits the peak inverse and maximum input voltages are halved, current output being as for half wave operation.



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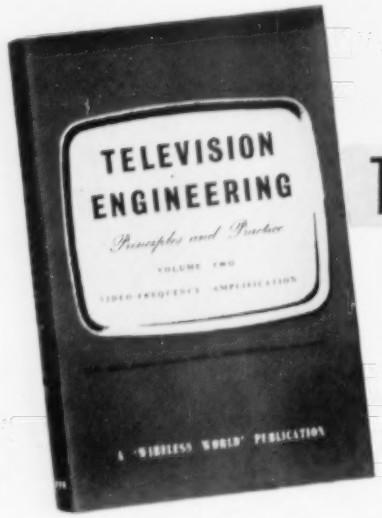

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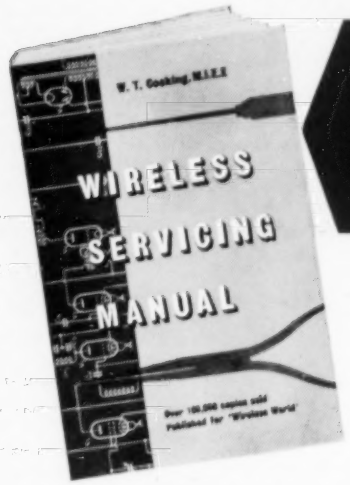

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has vacancies in the Electronics Division Extension at Bracknell, Berks

(i) **Experimental Officer** (Ref. 585/46) to assist in the design and development of electronic instruments for nuclear particle detecting equipments and to be responsible for development through to a production prototype.

Previous experience in the design of circuits for pulse and D.C. applications is required and some experience in the preparation of technical information for production is desirable.

(ii) **Experimental or Assistant Experimental Officer** (Ref. 584/46) to be responsible for the testing, calibration and inspection of electronic equipment. The duties will require ability to initiate design modifications for improved layout and reliability.

Previous experience in the use of electronic test equipment in pulse and D.C. applications is required. Some experience of the construction and maintenance of electronic equipment is desirable.

(iii) **Assistant Experimental Officer** (Ref. 586/46) to assist in the design and development of electronic instruments for nuclear particle detecting equipment.

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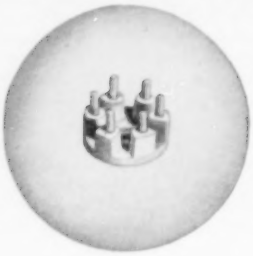

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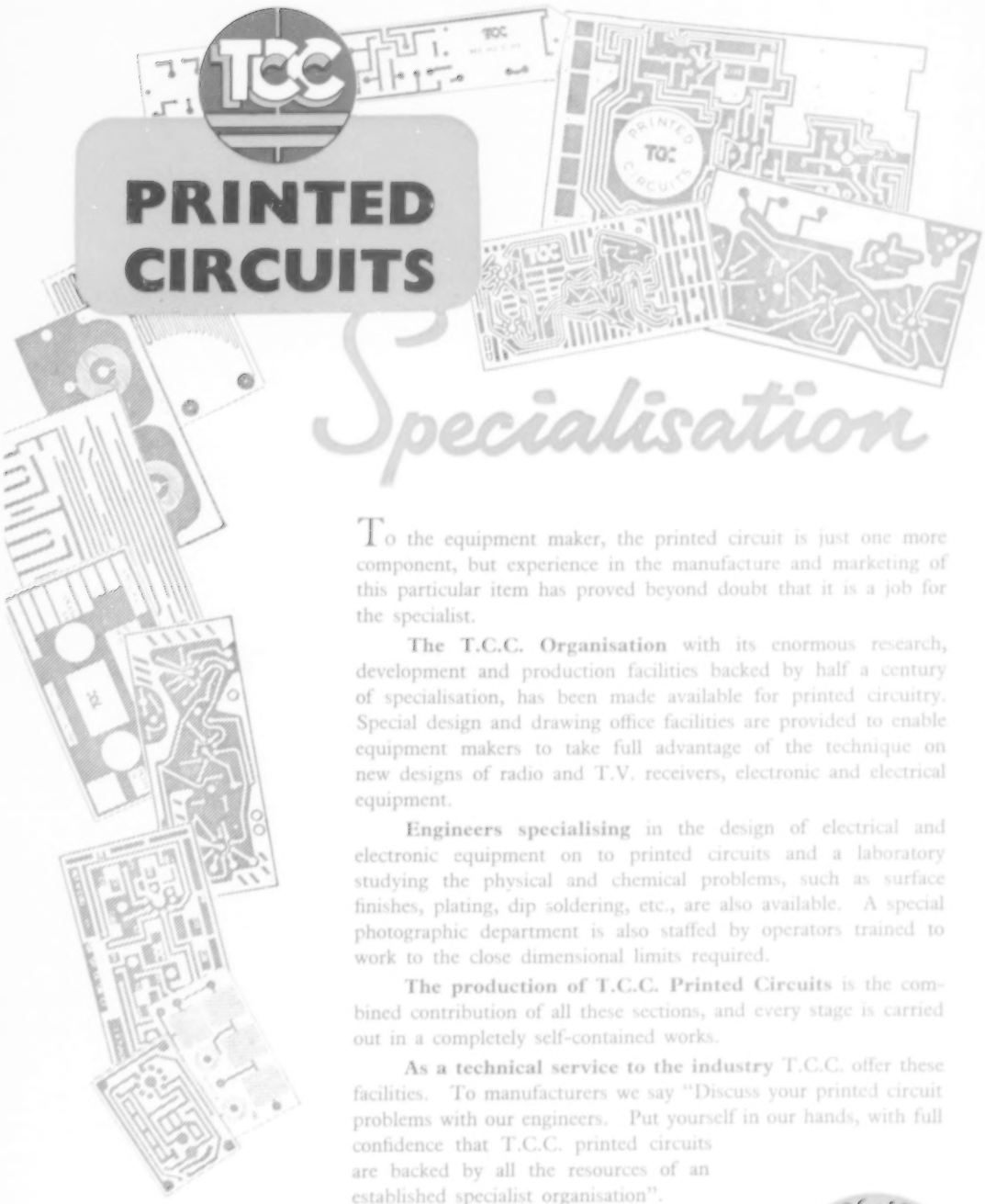

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The T.C.C. Organisation with its enormous research, development and production facilities backed by half a century of specialisation, has been made available for printed circuitry. Special design and drawing office facilities are provided to enable equipment makers to take full advantage of the technique on new designs of radio and T.V. receivers, electronic and electrical equipment.

Engineers specialising in the design of electrical and electronic equipment on to printed circuits and a laboratory studying the physical and chemical problems, such as surface finishes, plating, dip soldering, etc., are also available. A special photographic department is also staffed by operators trained to work to the close dimensional limits required.

The production of T.C.C. Printed Circuits is the combined contribution of all these sections, and every stage is carried out in a completely self-contained works.

As a technical service to the industry T.C.C. offer these facilities. To manufacturers we say "Discuss your printed circuit problems with our engineers. Put yourself in our hands, with full confidence that T.C.C. printed circuits are backed by all the resources of an established specialist organisation".

THE TELEGRAPH CONDENSER CO. LTD

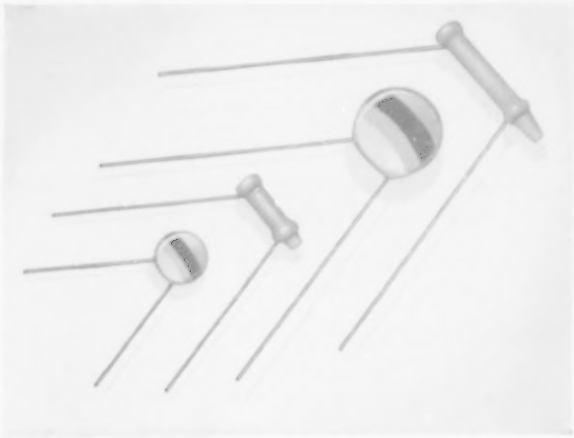
SPECIAL PRODUCTS DIVISION

NORTH ACTON · LONDON · W.3 · Telephone: ACORN 0061

1906 - 1956 GOLDEN JUBILEE

Tubular and Disc Ceramics

New!



Hunts announce their new ranges of Tubular and Disc Ceramics. Precise in their characteristics and robust in design, these capacitors are available in High-K, High-Q and Special Purpose Tubulars and in High-K Discs. Full technical specifications on request.

HIGH-Q TUBULAR
STANDARD CAPACITANCE RANGES

Capacitance Range Picofarads	Length of Insulated Tube Millimetres
*Type P100. Temperature Coefficient $(+100 \pm 100) \times 10^{-6}$ pF/pF/°C	
1.5 to 7	11
7.1 to 11	13
11.1 to 16	16
16.1 to 26	21
26.1 to 33	26
33.1 to 40	31
40.1 to 50	36
50.1 to 60	41
*Type N33. Temperature Coefficient $(-33 \pm 60) \times 10^{-6}$ pF/pF/°C	
5.0 to 27	11
27.1 to 45	13
45.1 to 69	16
69.1 to 100	21
100.1 to 140	26
140.1 to 180	31
180.1 to 220	36
220.1 to 250	41
*Type N750. Temperature Coefficient $(-750 \pm 250) \times 10^{-6}$ pF/pF/°C	
10.0 to 80	11
80.1 to 110	13
110.1 to 180	16
180.1 to 240	21
240.1 to 350	26
350.1 to 450	31
450.1 to 550	36
550.1 to 650	41

*Capacitors with closer limits of Temperature Coefficient can be supplied by special arrangement.

HIGH-K TUBULAR
STANDARD CAPACITANCE RANGE

Capacitance Picofarads	Length of Insulated Tube Millimetres	
	TB1000	TB3000
470	11	—
680	11	—
800	11	11
1000	13	11
1500	16	11
2000	19	11
2200	19	11
3000	—	16
3300	—	16
4000	—	19
4700	—	19
5000	—	21

POWER CERAMICS

Ranges of Power Ceramics and Temperature Compensating types are also available.
Details on application.

HIGH-K DISC
STANDARD CAPACITANCE RANGE

Capacitance Picofarads	Type Number	Diameter (over insulation) Millimetres
470	CD8K/2	9
680	CD8K/2	9
1000	CD9K/2	10.5
1500	CD9K/2	10.5
2200	CD11K/2	12
3300	CD12K/2	13
4700	CD14K/2	14.5

MATERIAL EMPLOYED

The temperature coefficient graph of the material is such that the capacitance increases with temperature from 20°C until it reaches a peak at approximately 50°C. With a further increase of temperature the capacitance falls, reaching its 20°C value again at approximately 80°C.
Maintenance of capacitance is thus ensured from 20°C to 80°C.

REGISTERED TRADE HUNTS MARK CAPACITORS

A. H. HUNT (Capacitors) LTD.

WANDSWORTH, LONDON, S.W.18. BAT 1083-7

And in Canada HUNT CAPACITORS (Canada) LTD. AJAX, ONTARIO